Mullard Technical Communications

APRIL 1972
Volume 12 Number 114

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Mullard Limited Mullard House Torrington Place London WC1E 7HD

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This journal records developments in the application of electronics in science, industry, and the home. It contains information on the properties and performance of solid-state devices, valves and tubes, and associated components and assemblies. The practical circuits given are designed to provide the best working conditions and performance for the devices used. The design principles may be used (so far as device ratings allow) for adaptation to meet particular requirements. Current data should be consulted before any circuit is considered for production. Mullard Limited can normally supply only those devices or assemblies listed in the Company's current catalogues. Availability in quantity should be confirmed.

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Editor: E. G. Evans
Central Technical Services, Telephone: 01-648 3471, Telex: 22194

Mullard Limited, Mullard House, Torrington Place, London, WC1E 7HD
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WIGHTMAN AND COMPANY LIMITED, 1/3 BRIXTON ROAD, LONDON SW9 6DS

Simple doppler radar using the CL8630 Gunn effect oscillator for the observation of small rotating objects

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Mullard Central Application Laboratory

UDC 621.396.962.23.029.64 [621-23/7]

A doppler radar using the Gunn effect oscillator CL8630 for measuring repetitive movements of a reflective object is described: the measurement of speed of rotation of a plastics propeller blade is used as a detailed example.

INTRODUCTION

The doppler effect and its exploitation in radar systems to detect relative motion are now generally well known, and the development of inexpensive solid state microwave generators, typified by oscillators using Gunn effect semiconductors, has made it practicable to use simple doppler radar systems for a number of industrial applications. An illustration of the principle, used in an intruder detector has already been published (Ref. 1).

The present article describes an application of this type of radar for the observation of the motion of small mechanisms, in particular to rotating objects. The approach followed is general and might be applied to any individual mechanical arrangement. However, a demonstration of the observation and measurement of the speed of rotation of a small, two-bladed plastics propeller has been constructed and is described in some detail.

Application

The rotation of the plastics blade, in the model described in this article, can be quite easily measured with a wooden board 10mm thick interposed between the detector and the rotating blade. Thus the detector can be used in many places where optically opaque barriers must of necessity exist between detector and the object whose speed is being measured. For example, the rotation of a rotating metal blade inside a plastics tube may be used to estimate fluid flow rates for certain fluids. (Unfortunately, water is not an appropriate fluid for this application.) The movement of enclosed machinery can also be followed easily from the outside, if part of the enclosure is transparent to microwaves.

GENERAL PRINCIPLES

The doppler effect is the phenomenon observed when energy, transmitted as a wave motion, is incident on a (relatively) moving object, where it undergoes a frequency shift upon reflection. The instantaneous incident and reflected frequencies when applied to a mixer device will produce a difference frequency which will vary with the relative speed of the moving object.

The amplitude of the difference or doppler signal will depend on the amplitude of the reflected signal which for an object of a given size will depend on the distance of the object from the receiver, and on its attitude.

For any system using coherent radio waves to produce doppler signals, it is necessary to have a source of energy, a transmitting aerial, a receiving aerial, and a detector; the latter might consist of a mixer device and a local oscillator.

In Ref. 1, which describes a doppler shift intruder detector system using CL8630 Gunn oscillator, it was shown that it was possible to simplify the system considerably by making components serve multiple functions. The Gunn diode, as well as being the source of microwave power, can also act as the mixer device and local oscillator; one aerial can fulfil the dual function for the transmitted and received signals.

If an object is reciprocating relative to the receiver, as when the object is rotating on a fixed shaft, it is possible to measure the average speed of reciprocation or rotation by counting the number of times the doppler signals reach a maximum over a given period. In fact, the frequency of the doppler signal will vary from zero to a maximum returning to zero again due to the relative forward and reverse motion. Thus the difference frequencies observed at the mixer appear as a complex repetitive waveform, with an envelope which varies with the number of times the maximum doppler signal is received. By suitable filtering and amplification, an unambiguous periodic signal may be produced from the envelope, varying with the speed of rotation.

The foregoing remarks perhaps over-simplify the actual situation when applied to a rotating target such as the twin-bladed, varying pitch propeller used in the demonstration model. In this case, the reflected wave will be extremely complex. Cancellations and reinforcements will occur, and by varying the position of the microwave source relative to the propeller, multiples of the fundamental rotational velocity can be found. Nevertheless, by observing the output waveform from the mixer on an oscilloscope whilst setting-up the system, positional adjustments may be made to produce the required (maximised) fundamental waveform.

GUNN OSCILLATOR

The CL8630 oscillator was designed for operation in the frequency band 10·675 to 10·700GHz at a typical output power level of 8mW from a supply voltage of +7V. Its applications include all forms of miniature radar systems. The centre frequency is 10·69GHz and the frequency temperature coefficient is −0·4MHz/degCmax. There is also some variation of frequency with supply voltage (frequency pushing), a typical value being 1·5MHz/V.

Range

As previously mentioned, the Gunn diode is also used in this type of application as the mixer for the detection of doppler signals. In any system using the detection of doppler signals, the maximum usable range is largely dependent of signal-to-noise ratio. Whilst the noise contribution from the Gunn diode is considerably higher than would be found in a microwave mixer diode, the much lower system cost and simplicity recommend its use. As an example of usable range; it was found quite possible to measure the speed of a roof-mounted extractor fan at a distance of about 7 metres using an aerial with a gain of 14dB. The signal-to-noise ratio and consequently the range can be improved by using an aerial with a higher gain.

AERIAL

The choice of an aerial suitable for the model was governed by the following factors: compactness, narrowness of the radiation pattern coupled with a reasonably high gain. A narrow beamwidth is also important when measuring speeds of moving parts in order to avoid obtaining spurious unwanted signal due to the movement of surrounding objects. A form of dielectric aerial was found to fulfil all the requirements (Ref. 2).

This aerial is made from a rectangular bar of Perspex (Plexiglass) 155mm long with a vee shaped section removed from the end in the H plane for first order mode propagation. The surfaces of the plastics should be left reasonably smooth. The absolute gain is about 14dB with a half power beamwidth of 24°; the first minima being at about 28° from the normal in both the E and H planes. The first side lobe maxima were some 13dB and 20dB down in the E and H planes respectively. The v.s.w.r. is better than 1·15 at 10·7GHz and matching to the cavity is achieved simply by varying the depth of penetration into a short length of WG16 waveguide.

The focusing action of the aerial in the H plane may be understood by considering a single ray of a TE_{10} wavefront being propagated within the dielectric (see Fig. 1). As can be seen, the ray will undergo a series of reflections from the dielectric-to-air interface both at the outside and inner surface of the vee until the angle of incidence is below the critical value. In fact, all the energy will not be contained in the dielectric before the critical angle is reached. A small percentage of the energy incident at the dielectric-to-air interface will escape progressively from the dielectric as energy is propagated through it.

The effective area (Ref. 3) of this aerial has been calculated as being about 1600mm^2. The aerial dimensions are given in Fig. 2.

DEMONSTRATION MODEL

The model was designed to demonstrate the observation and measurement of the speed of a two-bladed plastics propeller, diameter 50mm, driven by a variable speed electric motor. Signal processing circuits were developed to make the display on a digital counter direct reading in rev/min.

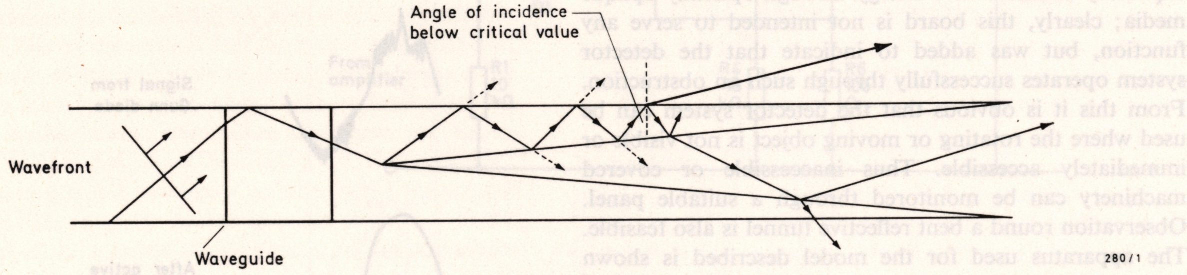
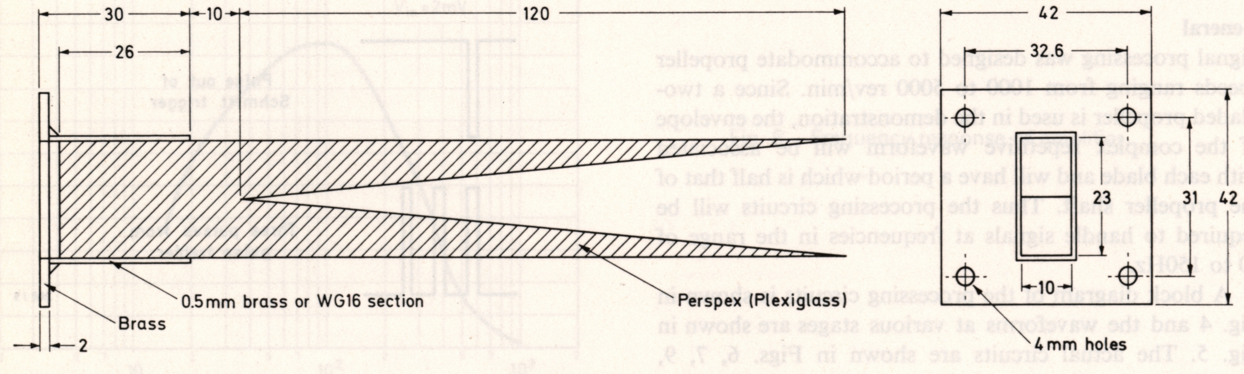


Fig. 1—Ray path diagram for aerial



All dimensions in mm

Fig. 2—Aerial and holder

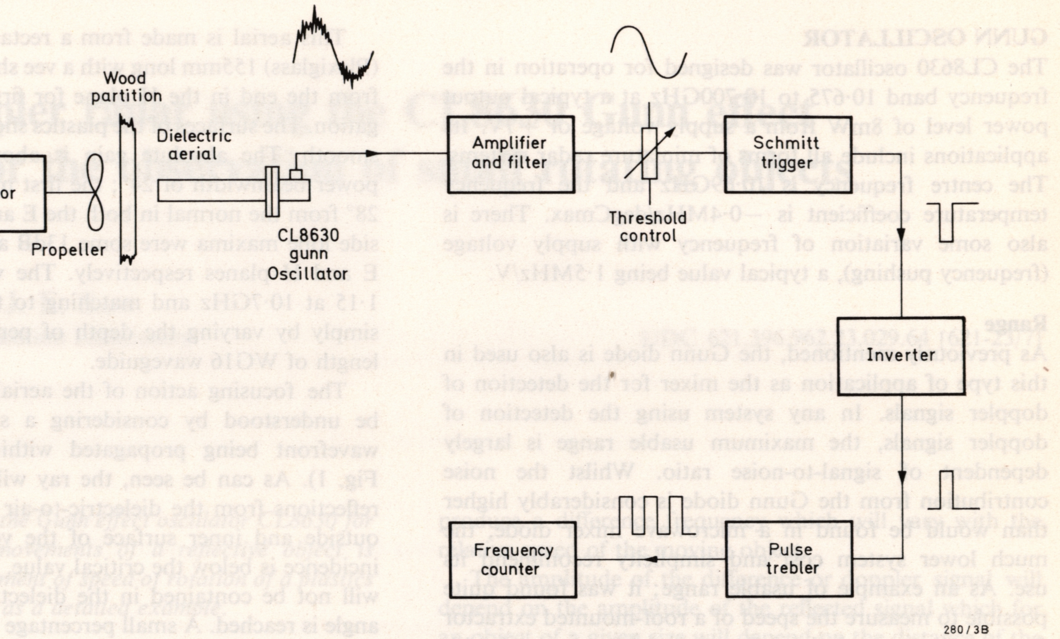

Wood partition
Dielectric aerial
Motor
Propeller
CL8630 gunn Oscillator
Motor speed controller
Amplifier and filter
Threshold control
Schmitt trigger
Inverter
Frequency counter
Pulse trebler
280/3B

Fig. 3—Schematic diagram of laboratory model showing typical waveforms

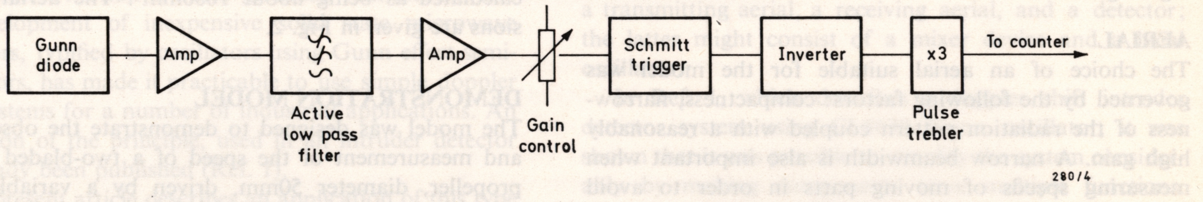

Gunn diode
Amp
Amp
Active low pass filter
Gain control
Schmitt trigger
Inverter
x3
Pulse trebler
To counter
280/4

Fig. 4—Block diagram of signal processing circuit

A wooden partition, 10mm thick, was put between the propeller and the aerial to demonstrate the penetration capability of microwave energy through optically opaque media; clearly, this board is not intended to serve any function, but was added to indicate that the detector system operates successfully through such an obstruction. From this it is obvious that the detector system can be used where the rotating or moving object is not visible or immediately accessible. Thus inaccessible or covered machinery can be monitored through a suitable panel. Observation round a bent reflective tunnel is also feasible. The apparatus used for the model described is shown schematically in Fig. 3.

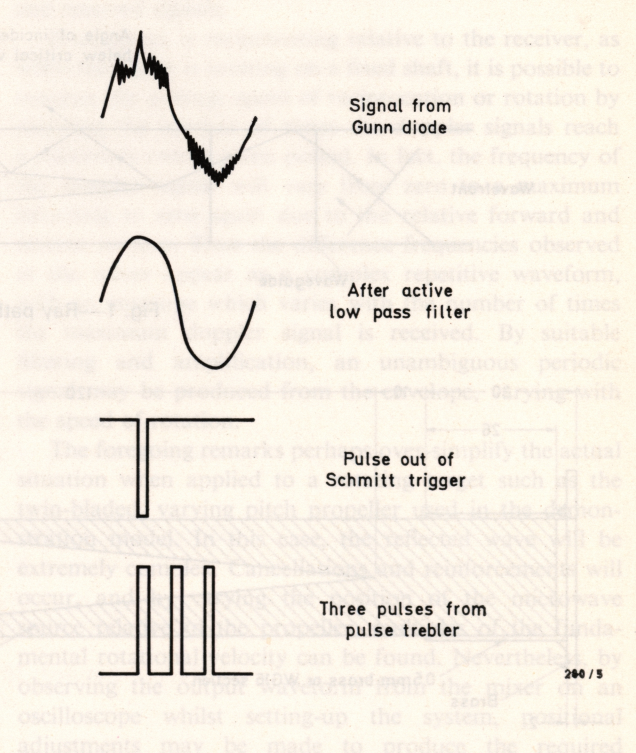


Fig. 5—Typical waveforms through signal processing circuit

ANCILLARY CIRCUITS

General

Signal processing was designed to accommodate propeller speeds ranging from 1000 to 5000 rev/min. Since a two-bladed propeller is used in the demonstration, the envelope of the complex repetitive waveform will be associated with each blade and will have a period which is half that of the propeller shaft. Thus the processing circuits will be required to handle signals at frequencies in the range of 30 to 150Hz.

A block diagram of the processing circuits is shown in Fig. 4 and the waveforms at various stages are shown in Fig. 5. The actual circuits are shown in Figs. 6, 7, 9, 10 and 11.

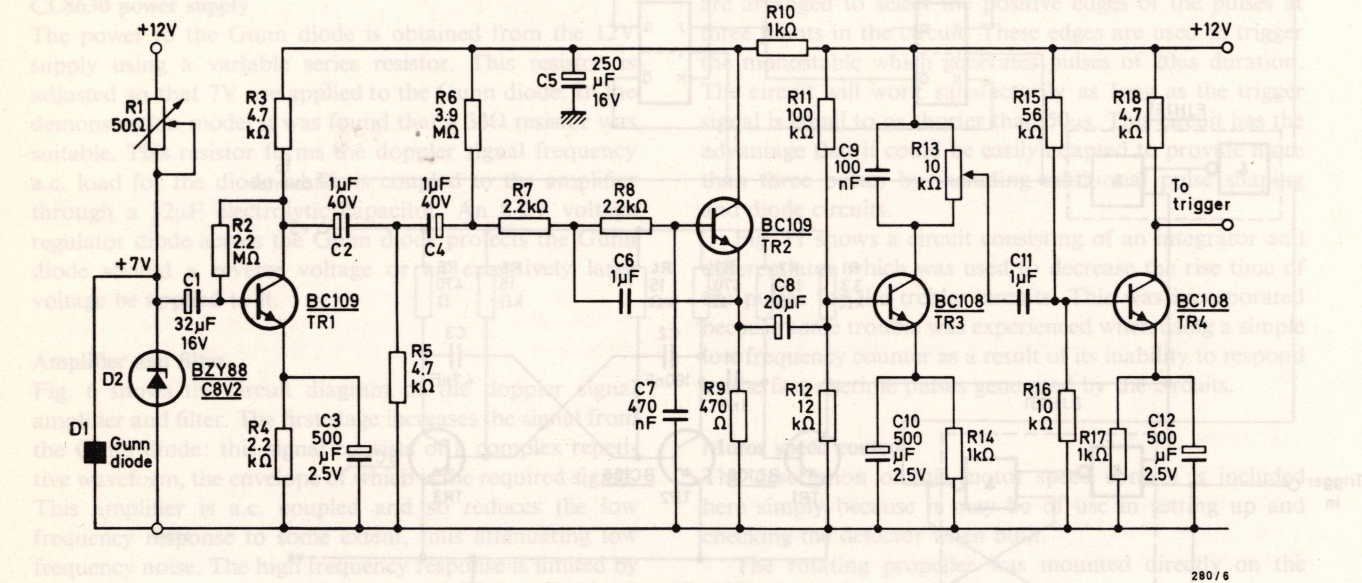

+12V
R1 50Ω
R3 4.7 kΩ
R2 2.2 MΩ
+7V
C1 32µF 16V
D2 BZY88 C8V2
D1 Gunn diode
BC109 TR1
R4 2.2 kΩ
C2 1µF 40V
C3 500 µF 2.5V
R6 3.9 MΩ
C4 1µF 40V
R5 4.7 kΩ
C5 250 µF 16V
R7 2.2kΩ
R8 2.2kΩ
C6 1µF
C7 470 nF
R9 470 Ω
BC109 TR2
R10 1kΩ
R11 100 kΩ
C9 100 nF
C8 20µF
R12 12 kΩ
R13 10 kΩ
BC108 TR3
C10 500 µF 2.5V
R14 1kΩ
R15 56 kΩ
R16 10 kΩ
C11 1µF
R17 1kΩ
R18 4.7 kΩ
BC108 TR4
C12 500 µF 2.5V
+12V
To trigger
280/6

Fig. 6—Amplifier filter

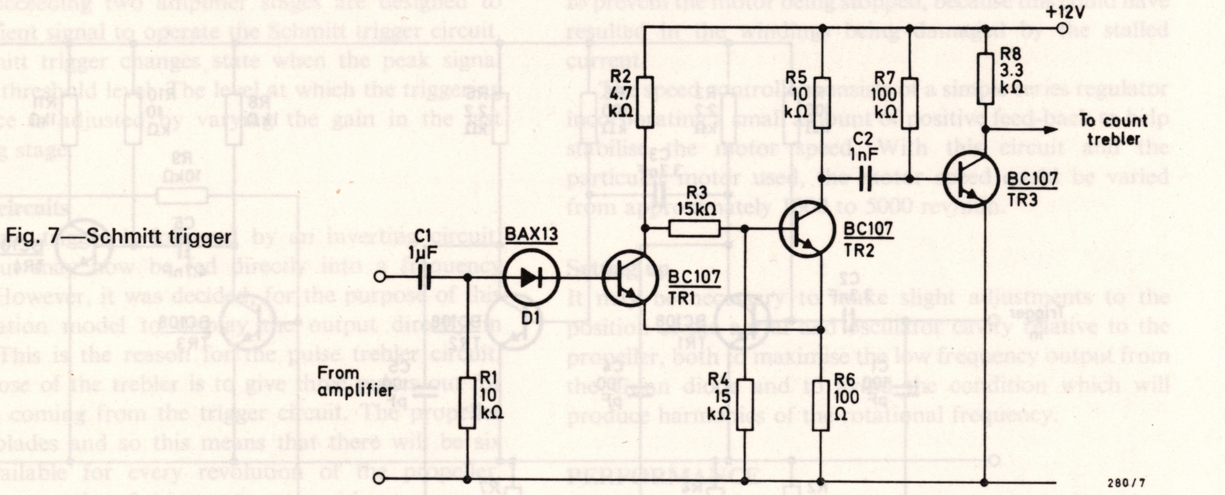

+12V
R2 4.7 kΩ
R5 10 kΩ
R7 100 kΩ
R8 3.3 kΩ
To count trebler
C2 1nF
BC107 TR3
R3 15kΩ
BC107 TR2
C1 1µF
BAX13
D1
BC107 TR1
From amplifier
R1 10 kΩ
R4 15 kΩ
R6 100 Ω
280/7

Fig. 7—Schmitt trigger

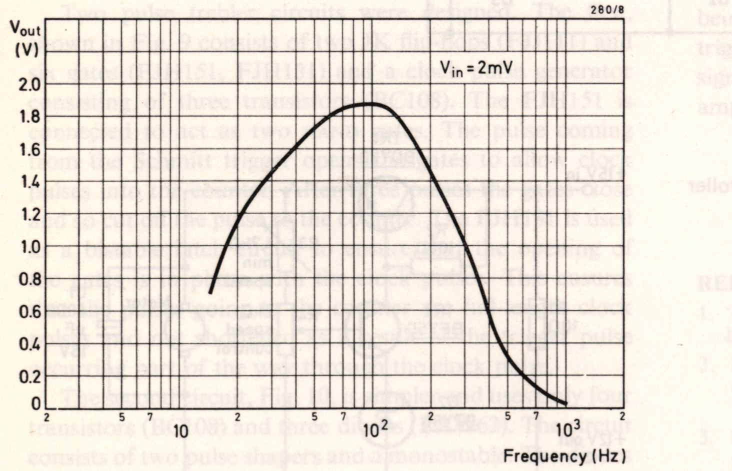


Fig. 8—Frequency response of amplifier

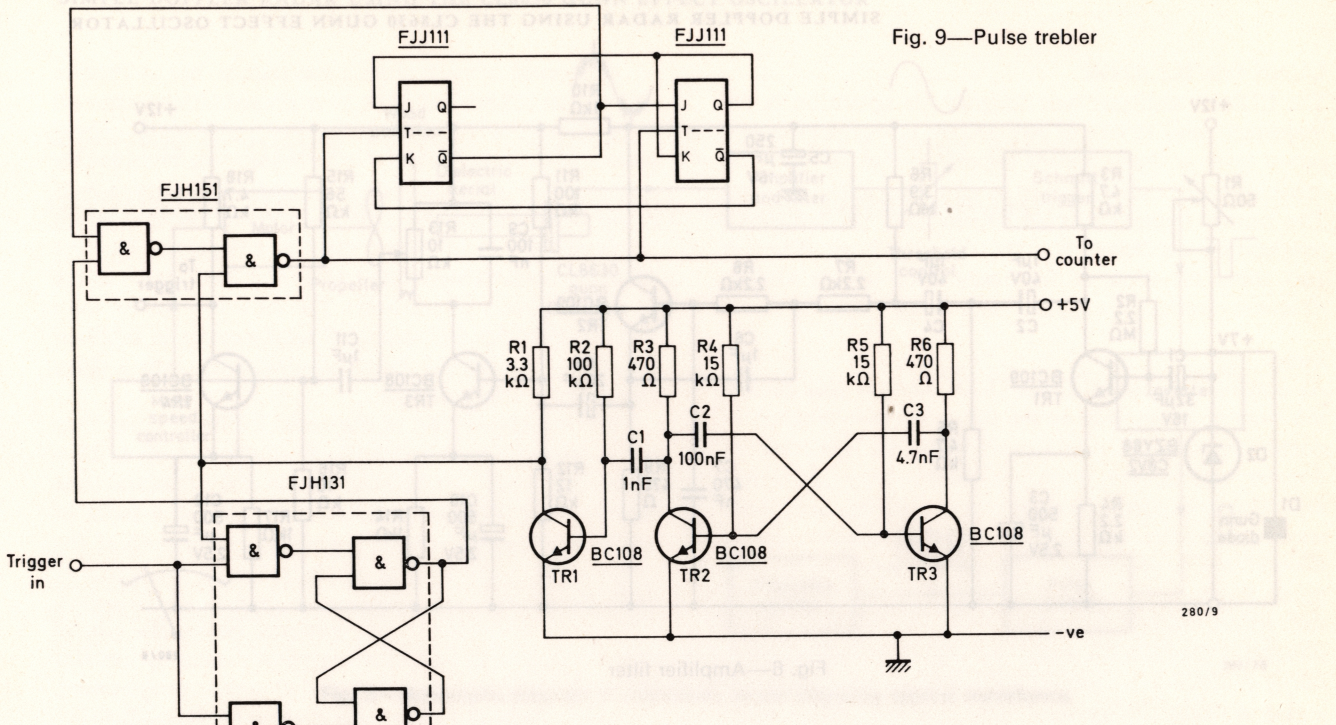


Fig. 9—Pulse trebler

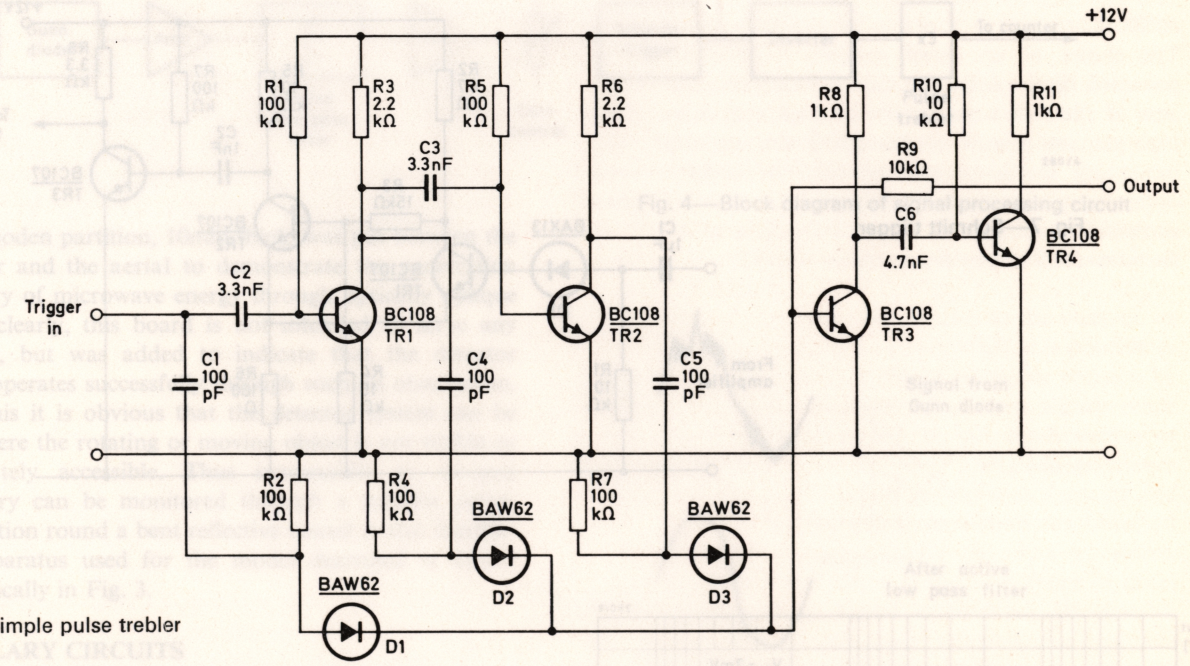


Fig. 10—Simple pulse trebler

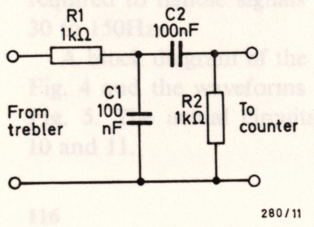


Fig. 11 (above)—Pulse shaper

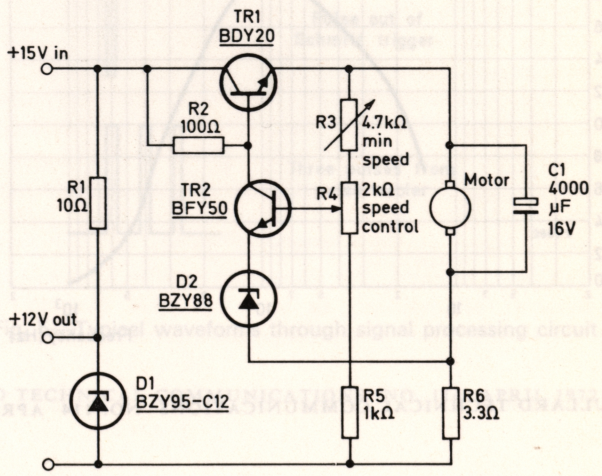


Fig. 12 (right)—Speed controller

CL8630 power supply

The power to the Gunn diode is obtained from the 12V supply using a variable series resistor. This resistor is adjusted so that 7V are applied to the Gunn diode. In the demonstration model it was found that a 33Ω resistor was suitable. This resistor forms the doppler signal frequency a.c. load for the diode which is coupled to the amplifier through a 32μF electrolytic capacitor. An 8·2V voltage regulator diode across the Gunn diode protects the Gunn diode should a reverse voltage or an excessively large voltage be applied to it.

Amplifier and filter

Fig. 6 shows the circuit diagram of the doppler signal amplifier and filter. The first stage increases the signal from the Gunn diode: this signal consists of a complex repetitive waveform, the envelope of which is the required signal. This amplifier is a.c. coupled and so reduces the low frequency response to some extent, thus attenuating low frequency noise. The high frequency response is limited by a single stage active filter which has a cut-off frequency of 150Hz and falls off at 12dB per octave above that frequency. Fig. 8 shows the frequency response of the amplifier.

The succeeding two amplifier stages are designed to give sufficient signal to operate the Schmitt trigger circuit. The Schmitt trigger changes state when the peak signal reaches a threshold level. The level at which the triggering takes place is adjusted by varying the gain in the last amplifying stage.

Counting circuits

The Schmitt trigger is followed by an inverting circuit. The output may now be fed directly into a frequency counter. However, it was decided, for the purpose of this demonstration model to display the output directly in rev/min. This is the reason for the pulse trebler circuit. The purpose of the trebler is to give three pulses out for every one coming from the trigger circuit. The propeller has two blades and so this means that there will be six pulses available for every revolution of the propeller. These pulses are then fed into a counter with a ten second gate time. The display on the counter is then a direct indication of the number of revolutions per minute of the propeller.

Two pulse trebler circuits were designed. The first, shown in Fig. 9 consists of two JK flip-flops (FJJ111) and six gates (FJH151, FJH131) and a clock pulse generator consisting of three transistors (BC108). The FJH151 is connected to act as two NAND gates. The pulse coming from the Schmitt trigger opens the gates to allow clock pulses into the counter. After three pulses the gates close and so cut off the pulse to the counter. The FJH131 is used as a bistable latch circuit to ensure that the opening of the gates is in phase with the clock pulses. This ensures that the pulses going to the counter are full-length clock pulses and not shortened as a result of the trigger pulse occurring part of the way through the clock pulse.

The second circuit, Fig. 10, is simpler and uses only four transistors (BC108) and three diodes (BAW62). The circuit consists of two pulse shapers and a monostable. The diodes are arranged to select the positive edges of the pulses at three points in the circuit. These edges are used to trigger the monostable which generates pulses of 20μs duration. The circuit will work satisfactorily as long as the trigger signal is equal to or shorter than 50μs. This circuit has the advantage that it could be easily adapted to provide more than three pulses by including additional pulse shaping and diode circuits.

Fig. 11 shows a circuit consisting of an integrator and differentiator which was used to decrease the rise time of the pulses for the trebler circuits. This was incorporated because some trouble was experienced when using a simple low frequency counter as a result of its inability to respond to the fast risetime pulses generated by the circuits.

Motor speed control

The description of this motor speed control is included here simply because it may be of use in setting up and checking the detector when built.

The rotating propeller was mounted directly on the shaft of a 24V d.c. parallel wound motor. This motor was operated from the control circuit shown in Fig. 12. This enabled the motor speed to be varied by means of a potentiometer. A preset potentiometer was incorporated to prevent the motor being stopped, because this could have resulted in the windings being damaged by the stalled current.

The speed controller consists of a simple series regulator incorporating a small amount of positive feed-back to help stabilise the motor speed. With this circuit and the particular motor used, the motor speed could be varied from approximately 1000 to 5000 rev/min.

Setting up

It may be necessary to make slight adjustments to the position of the aerial and oscillator cavity relative to the propeller, both to maximise the low frequency output from the Gunn diode and to avoid the condition which will produce harmonics of the rotational frequency.

PERFORMANCE

With the CL8630 oscillator and the dielectric aerial approximately 150mm from the propeller, signals varying between 0·5V p–p and 2V p–p were obtained at the output of the amplifier. The variation was due to the aerial being aimed at different parts of the propeller. The Schmitt trigger circuit worked satisfactorily over the range of signals available. The trigger circuit gave pulses of 12V amplitude lasting 50μs.

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Design of a microwave oven cavity using two magnetrons, type YJ1371

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UDC 643.336:621.385.64.029.63

A 3kW two-plate microwave oven cavity is described which is energised by two 1·5kW magnetrons, type YJ1371. Excitation by one magnetron only allows convenient operation at half power.

INTRODUCTION

A microwave oven is a closed cavity dimensioned to resonate at several frequencies in the industrial microwave band (2·450±0·025 GHz). When microwave energy at a single resonant frequency is coupled into the cavity, a standing wave pattern is produced with characteristic areas of low and high field intensity, nodes and antinodes. To excite all the resonant modes in the cavity, a mode changer is incorporated to pull the frequency of the microwave generator over a small range. This causes the standing wave patterns to overlap resulting in enlarged areas of high intensity which can be used to promote even heating of a load.

This article describes the design of a 3kW two-plate size cavity energised by two magnetrons, type YJ1371. The cavity is suitable for operation at a half power setting with one magnetron.

CAVITY DESIGN

When designing a two-magnetron cavity, it is important to ensure that, in addition to the usual microwave oven requirements for a large number of resonant modes with a suitable distribution, the intercoupling between the two magnetrons is a minimum. That is, the power reflected back into each magnetron under worst-case conditions must be less than the permissible maximum derived from the published data. This reflected power consists of two components, one from each magnetron and hence not necessarily at the same frequency.

Intercoupling can be reduced to a minimum by designing the cavity so that it has two independent sets of resonances. That is, the nodes of each set must coincide as far as possible with the antinodes of the other and vice versa. Strictly, this condition will only hold for an empty cavity since any load present will inevitably perturb the field. However, intercoupling between the sets of resonances will still be low even with a load present. This is because small loads, although generally giving high reflected powers, have little effect on the field distribution, while large loads which strongly perturb the field do not result in much reflection.

In general, a mode will be available for each resonant frequency given by:

$$1/f_{res} = \tfrac{1}{2}[(l/a_x)^2+(m/a_y)^2+(n/a_z)^2]^{\frac{1}{2}},$$

where a_x, a_y, and a_z are the dimensions of the cavity in the directions x, y, and z, and l, m, and n are integers. Each set of integers represents a resonant mode.

The dimensions of a suitable cavity to operate in the frequency range 2·425 to 2·475GHz were selected with the aid of a computer. These, and the resonant modes, are given in Table 1.

TABLE 1

Dimensions of selected cavity and its resonant modes

Length 505mm Depth 315mm Height 280mm

	Mode numbers			*Resonant frequency*
	l	*m*	*n*	*(GHz)*
Set 1	6	1	3	2·445
	4	3	3	2·445
	4	0	4	2·448
Set 2	1	5	1	2·457
	2	0	5	2·452
	8	0	1	2·434
	5	4	1	2·472
	0	5	1	2·439

Mode charts for the rear and top faces of the cavity together with the magnetron launch positions are given in Figs. 1 and 2. It is seen that the first three modes (set 1) in Table 1 can be directly excited from the rear launch port, and the fourth (set 2), from the top launch port. Furthermore, neither set of modes excites the other. The other modes in Table 1, although not directly excited, are likely to be excited when the mode changers are operating.

Magnetron launch system

The usual type of magnetron launch system consists of a length of rectangular waveguide which couples the magnetron to the cavity. However, this system is impracticable for the YJ1371 magnetron because the diameter of the cooling assembly is greater than the broad dimension of the waveguide. This would make the ducting required to feed the magnetron cooling air into the oven cavity to remove cooking vapours somewhat complex. Therefore, to overcome this problem, a small auxiliary cavity is used as the magnetron mount (Fig. 3).

In Fig. 3, the magnetron cooling assembly fits into a skirt extending from the auxiliary cavity. The top face of the cavity is perforated to allow the cooling air, flowing coaxially with the output probe, to pass via the output waveguide, into the oven cavity. The input waveguide extends radially from the auxiliary cavity and then bends through 90° to form the inlet port to the main cavity.

The diameter of the auxiliary cavity is chosen to be the same as that of a circular waveguide dimensioned to operate close to cut-off at microwave cooking frequencies. Thus, the cavity will resonate in the TM_{010} mode at a frequency independent of its axial length, and the transition from magnetron to output waveguide will be broadband.

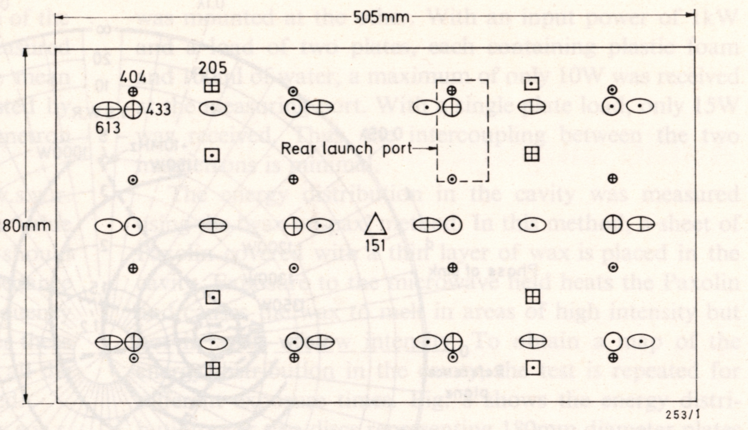

Fig. 1—Mode chart for rear face showing antinodes of all modes and position of rear launch port

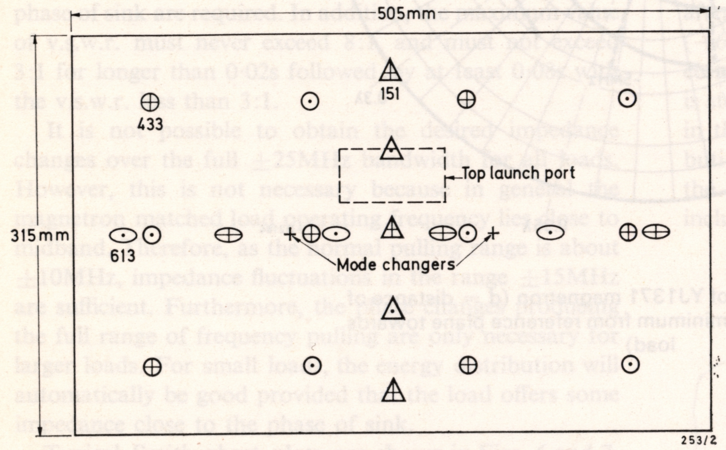

Fig. 2—Mode chart for top face showing antinodes of all modes and positions of top launch port and mode changers

To determine the optimum launch positions for a range of loads, a prototype cavity was constructed on which the sides carrying the launch mounts were slightly longer than the basic cavity dimensions. This allowed the positions of the launch ports to be varied over a small range about the theoretically correct positions.

Mode Changers

As in most microwave ovens, a mode changer system is installed to pull the magnetron operating frequency and to excite modes not directly excited from the launch positions. In this way a more even field distribution is achieved. Twin single-bladed mode changers are used, and these rotate at different speeds (30 and 60 rev/min) to ensure that their orientation with respect to each other is not critical. One of the blades is shown in Fig. 4.

PERFORMANCE

Cold measurements

The impedance offered to each magnetron by the cavity was measured with a swept-frequency Smith chart plotter. The initial measurements were concerned with optimising the launch positions as described in a previous section. To achieve this objective, the maximum amplitude of the

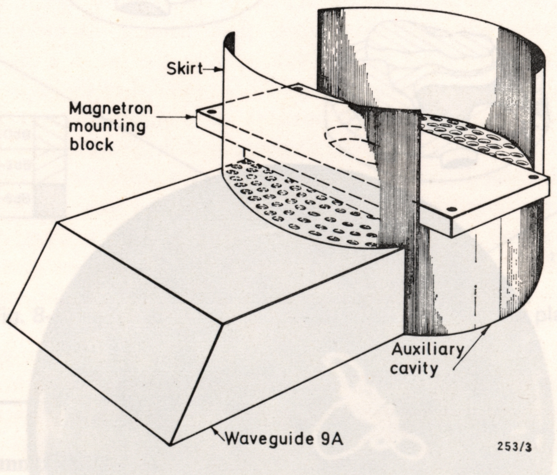

Fig. 3—Magnetron launch mount

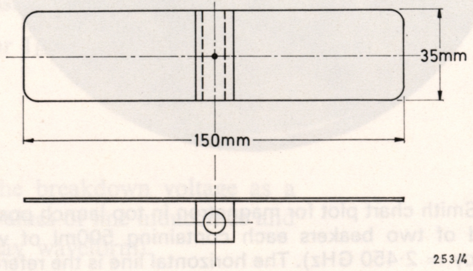

Fig. 4—Mode changer blade

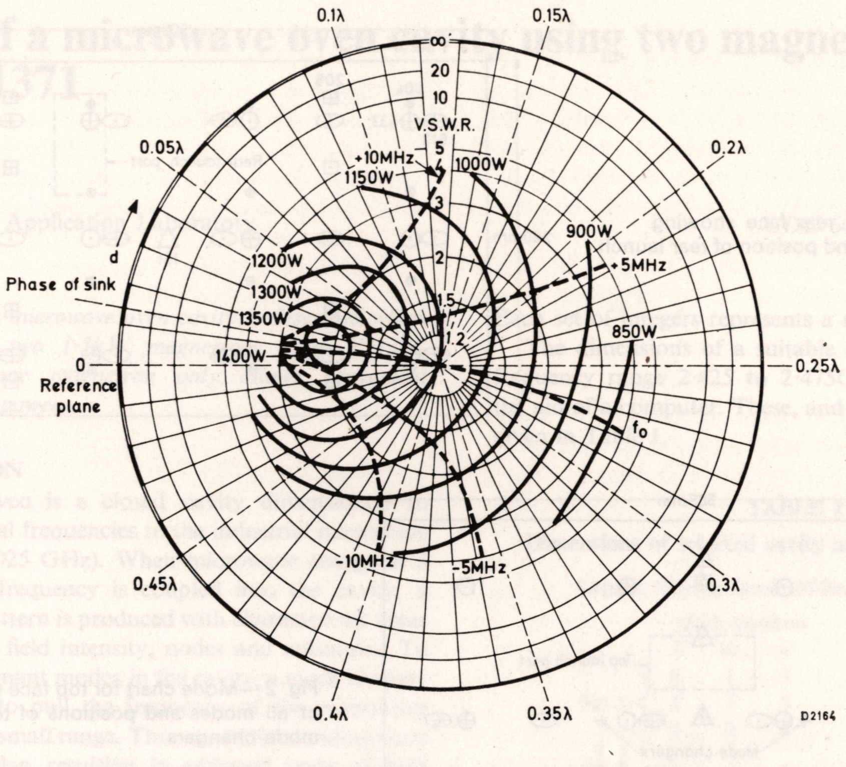

Fig. 5—Rieke diagram of YJ1371 magnetron (d = distance of voltage standing wave minimum from reference plane towards load)

Fig. 6—Smith chart plot for magnetron in top launch position and load of two beakers each containing 500ml of water (frequency = 2·450 GHz). The horizontal line is the reference plane (see Fig. 5)

Fig. 7—Smith chart plot for magnetron in rear launch position and load of two plates each containing polyeurathene foam and 100ml of water (frequency = 2·450 GHz). The horizontal line is the reference plane (see Fig. 5)

v.s.w.r. and also the change in amplitude and phase of the v.s.w.r. with rotation of the mode changers were measured for a variety of loads. The absolute value of the mean phase was not measured since this could be adjusted by altering the length of waveguide between the magnetron and the cavity inlet port.

It is required that the mean amplitude of the v.s.w.r. should fluctuate about the maximum continuous value quoted in the published data, and that the phase should fluctuate about the phase of sink. That is, the impedance plot should cut as many contours of constant frequency on the magnetron Rieke diagram as possible. Under these conditions, maximum frequency pulling occurs and all the required resonances in the cavity should be excited.

For the YJ1371 magnetron (Fig. 5), a mean v.s.w.r. value of 2·5:1 and a phase change of ±0·05λ about the phase of sink are required. In addition, the maximum value of v.s.w.r. must never exceed 8:1, and must not exceed 3:1 for longer than 0·02s followed by at least 0·08s with the v.s.w.r. less than 3:1.

It is not possible to obtain the desired impedance changes over the full ±25MHz bandwidth for all loads. However, this is not necessary because in general the magnetron matched load operating frequency lies close to midband. Therefore, as the normal pulling range is about ±10MHz, impedance fluctuations in the range ±15MHz are sufficient. Furthermore, the phase changes producing the full range of frequency pulling are only necessary for larger loads. For small loads, the energy distribution will automatically be good provided that the load offers some impedance close to the phase of sink.

Typical Smith chart plots are shown in Figs. 6 and 7. When compared with Fig. 5, it is seen that in both cases the mean v.s.w.r. is about 2:1 and that the phase varies about the phase of sink.

Hot measurements

Each magnetron was set to deliver at least 1·5kW into a standard load of two beakers each containing 500ml of water. With both magnetrons operating, more than 3kW load power was measured.

To determine the degree of intercoupling between the launch positions, a magnetron was mounted at one launch port and a measuring probe connected to a power meter was mounted at the other. With an input power of 1kW and a load of two plates, each containing plastic foam and 100ml of water, a maximum of only 10W was received at the measuring port. With a single plate load, only 15W was received. Thus, the intercoupling between the two magnetrons is minimal.

The energy distribution in the cavity was measured using the 'wax-on-pax' method. In this method, a sheet of Paxolin covered with a thin layer of wax is placed in the cavity. Exposure to the microwave field heats the Paxolin and causes the wax to melt in areas of high intensity but not in areas of low intensity. To obtain a map of the energy distribution in the cavity, the test is repeated for different exposure times. Fig. 8 shows the energy distribution over two discs representing 180mm diameter plates placed 40mm up from the floor of the cavity. Areas melting after x seconds (0dB), 2x seconds (−3dB) and 4x seconds (−6dB) are shown. Although the right-hand disc shows a certain amount of energy variation, the thermal gradient is steep and this should result in good thermal conduction in the load. Efforts to further improve the energy distribution are continuing and work is being carried out on the design of the mode changers and on the effect of including a plastic liner in the cavity.

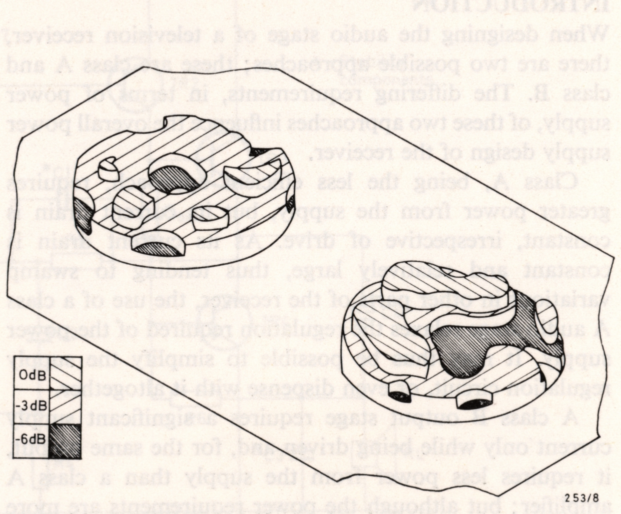

Fig. 8—Energy distribution over two 180mm diameter plates, 40mm above floor of cavity

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Mullard Technical Communications

Vol. 12, No. 113, January 1972, pages 94 to 111

Transient voltage suppression using transient suppressor diodes

In this article, on page 98, the expression for I_{RSM} should read:

$$I_{RSM} = \frac{V_{SM}}{Z_S}$$

Also, the caption to Fig. 6 should read:

Fig. 6—Normalised excess voltage above the breakdown voltage as a function of dynamic impedance for various values of line inductance and transient duration (step-and-exponential-decay waveform).

On page 106, the ordinate to Fig. 24 is P_{RSM}(W).

Audio amplifiers giving up to 3W for use with TBA750

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UDC 621.375.1.029.4.026.441(TBA750)

This article discusses class A and class B designs of tv audio stages for use with the TBA750 intercarrier-sound integrated circuit. General and detailed circuits are provided for class B, and a detailed class A circuit is also given. Reference is also made to possible transformer-coupled class A circuits.

INTRODUCTION

When designing the audio stage of a television receiver, there are two possible approaches; these are class A and class B. The differing requirements, in terms of power supply, of these two approaches influence the overall power supply design of the receiver.

Class A, being the less efficient approach, requires greater power from the supply, but its current drain is constant, irrespective of drive. As its current drain is constant and relatively large, thus tending to swamp variations in other parts of the receiver, the use of a class A audio stage relaxes the regulation required of the power supply. It may thus be possible to simplify the supply regulation circuit, or even dispense with it altogether.

A class B output stage requires a significant supply current only while being driven and, for the same output, it requires less power from the supply than a class A amplifier; but although the power requirements are more modest, the supply regulation must be good to withstand the variable loading imposed upon it by the amplifier.

The purpose of this article is to set out a number of specific designs of both class A and B approach together with tentative suggestions for other designs, for use with the TBA750 intercarrier-sound integrated circuit.

CLASS B DESIGNS

Two designs using the class B approach are discussed below; these are a three-transistor circuit and a four-transistor circuit.

Three-transistor circuit

Fig. 1 shows the three-transistor circuit. It is a conventional d.c. coupled arrangement with the input capacitively coupled to pin 16 of the TBA750.

The mid point is set up by R_1, R_2, and R_5. The resistor R_5 is incorporated to reduce the effects of V_{be} spread of TR_1 on the mid point; without this, a preset resistor would be needed to set up the mid-point. A stabistor (two element stabilising diode) and a preset resistor are included for setting and stabilising the quiescent output current; it is possible to use instead, a transistor or thermistor arrangement.

The source impedance of the integrated circuit at pin 16 is approximately 4·7kΩ and, by suitable choice of R_1 (the feedback resistor), the circuit can be made to have the correct sensitivity to be driven direct from this source. (The unloaded voltage swing at pin 16 is $\geqslant$ 1·0V r.m.s. for 15kHz deviation).

Four-transistor circuit

A four-transistor circuit, driven from pin 15 of the integrated circuit is shown in Fig. 2. The circuit is again a conventional d.c. coupled type, except for the method of connecting the input to the integrated circuit. Use is made of the well-defined direct output voltage at pin 15 of the integrated circuit to set up the mid-point of the amplifier, thus saving a capacitor. The resistors R_1 and R_2 determine the mid-point voltage and R_3 provides quiescent current for the output emitter follower of the integrated circuit.

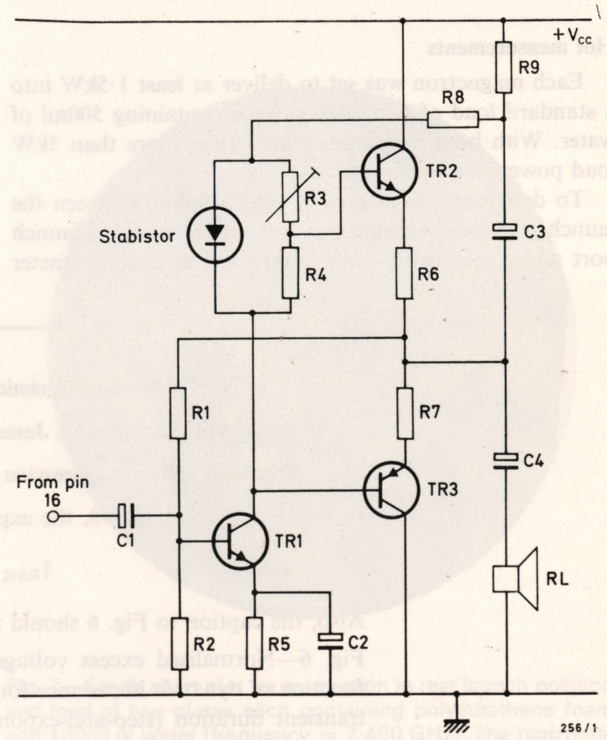

Fig. 1—Class-B circuit using three transistors

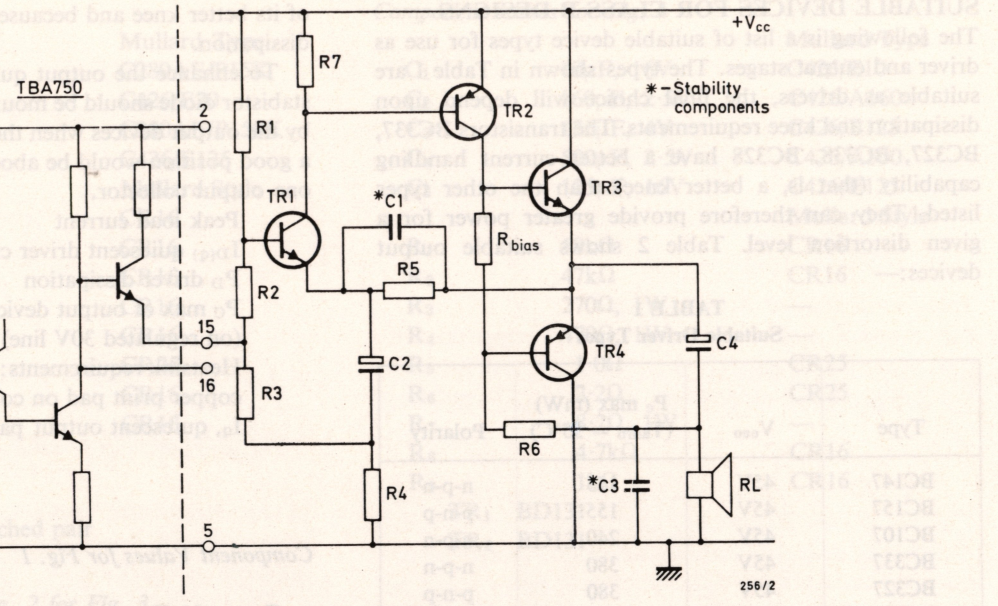

Fig. 2—Class-B circuit using four transistors

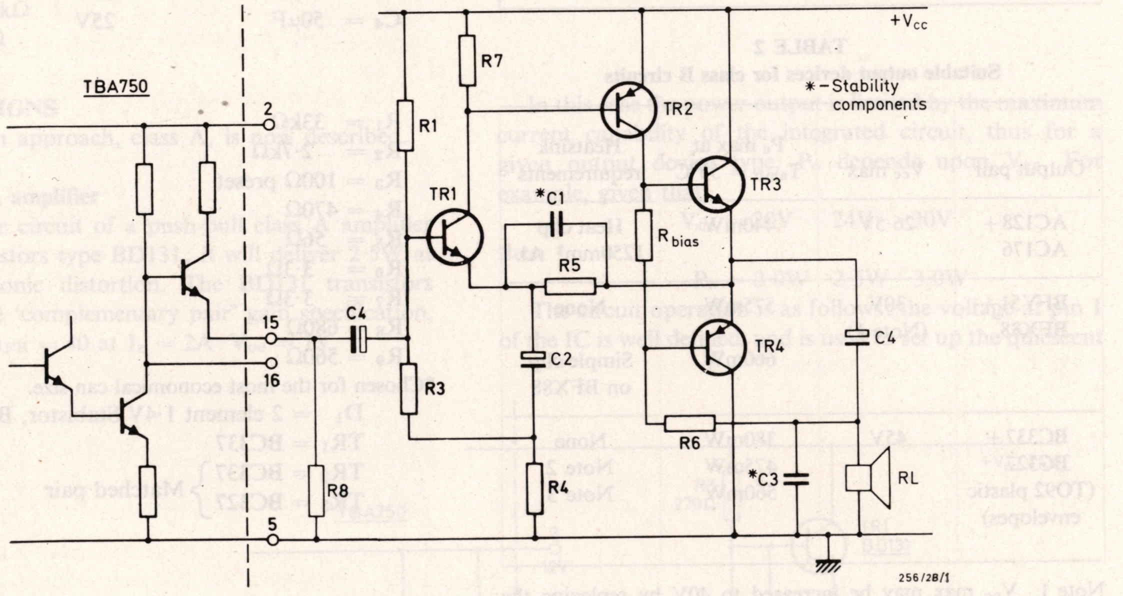

Fig. 3—Alternative class-B circuit using four transistors

If there is a significant amount of ripple on the supply line the circuit shown in Fig. 2 is not directly suitable because R_1 will inject the ripple on to the input. To overcome this problem, the modified circuit of Fig. 3 should be used. Here the mid-point is set up by the normal bias chain. The low output impedance of the TBA750 (100Ω) decouples the ripple current because of R_1: the value of C_1 should be sufficient to ensure adequate decoupling by this means.

The open-loop gain of the amplifier is large, and therefore a large amount of feedback can be applied; a high open-loop distortion can thus be tolerated. A way of using this to advantage is to simplify the bias arrangements of the output pair, that is, to use a simple resistor bias, as shown. There is an added advantage in this approach because R_{bias} must be chosen such that under worst case conditions the output pair is just beginning to pass some quiescent current, and this means that the dissipation in the output pair approaches the ideal class B case; heatsink requirements are thus reduced.

SUITABLE DEVICES FOR CLASS B DESIGNS

The following is a list of suitable device types for use as driver and output stages. The types shown in Table 1 are suitable as drivers, the final choice will depend upon dissipation and knee requirements. The transistors BC337, BC327, BC338, BC328 have a better current handling capability (that is, a better knee) than the other types listed. They can therefore provide greater power for a given distortion level. Table 2 shows suitable output devices:—

TABLE 1
Suitable Driver Types

Type	V_{ceo}	P_c max (mW) ($T_{amb} = 55°C$)	Polarity
BC147	45V	155	n-p-n
BC157	45V	155	p-n-p
BC107	45V	240	n-p-n
BC337	45V	380	n-p-n
BC327	45V	380	p-n-p
BC338	25V	380	n-p-n
BC328	25V	380	p-n-p

TABLE 2
Suitable output devices for class B circuits

Output pair	V_{cc} max	P_c max at $T_{amb} = 55°C$	Heatsink requirements
AC128+ AC176	26·5V	440mW	Heat clip 1250mm^2 Al.
BFY51+ BFX88	30V (Note 1)	575mW	None
		660mW	Simple clip on BFX88
BC337+ BC327 (TO92 plastic envelopes)	45V	380mW	None
		475mW	Note 2
		560mW	Note 3

Note 1. V_{cc} max may be increased to 40V by replacing the BFY51 with a BFX84.

Note 2. Device soldered to printed circuit (lead length ≮ 3mm) with 100mm^2 pad of Cu. print on collector lead.

Note 3. As Note 2 but with small heat clip in addition.

SPECIFIC CLASS B EXAMPLES

Two examples of a specific design, using the circuits already mentioned are now given.

Three-transistor circuit

Design details:

V_{cc} 30V
P_o 1·9W (to clipping)
R_L 40Ω

Devices and operating conditions

BC337 and BC327 were chosen as the output pair with a BC337 as driver. This device was chosen as driver because of its better knee and because it will handle the necessary dissipation.

To enhance the output quiescent current stability, the stabistor diode should be mounted where it will be warmed by the output devices when they become hot; for example, a good position would be above the 100mm^2 print pad on one output collector.

Peak load current = 310mA
$I_{D(q)}$ quiescent driver current = 11·5mA
P_D driver dissipation = 175mW
P_C max of output devices
(on regulated 30V line) = 560mW
Heatsink requirements: small heat clip and 100mm^2 copper print pad on collector lead.
I_q, quiescent output pair current = 2 to 3mA

Component Values for Fig. 1

			Mullard Type
C_1	= 64μF	10V*	C426/D64
C_2	= 320μF	2·5V	C426/A320
C_3	= 25μF	25V*	C426/F25
C_4	= 50μF	25V	C426/F50
			Mullard Style
R_1	= 33kΩ		CR16
R_2	= 2·7kΩ		CR16
R_3	= 100Ω preset		—
R_4	= 470Ω		CR16
R_5	= 56Ω		CR16
R_6	= 3·3Ω		CR16
R_7	= 3·3Ω		CR16
R_8	= 680Ω		CR16
R_9	= 560Ω		CR16

*Chosen for the most economical can size.

D_1 = 2 element 1·4V Stabistor, BZX75—C1V4
TR_1 = BC337
TR_2 = BC337 } Matched pair
TR_3 = BC327 }

Four-transistor design

Design details:

$V_{cc} = 24V$
P_o = 2·5W (to clipping)
$R_L = 25Ω$

Devices and operating conditions

Output pair: BFY51 and BFX88
Driver : BC328
Pre-driver : BC148
Peak load current, I_L = 450mA
Quiescent driver current $I_{D(q)}$ = 30mA
Driver dissipation P_D = 280mW
P_C max of output devices
(on 24V regulated line) = 580mW
Heatsinks: the BFX88 should have a small heat clip.

Component values for Fig. 2

		Mullard Type
C_1	= 15nF	C280AE/P15K
C_2	= 20μF, 16V	C426/E20
C_3	= 150nF	C280AE/P150K
C_4	= 125μF, 16V	C426/E125
		Mullard Style
R_1	= 22kΩ	CR16
R_2	= 12kΩ	CR16
R_3	= 3·9kΩ	CR16
R_4	= 120Ω	CR16
R_5	= 1KΩ	CR16
R_6	= 390Ω	CR25
R_7	= 4·7kΩ	CR16
R_{bias}	= 15Ω	CR16

TR_1 BC148
TR_2 BC328
TR_3 BFY51 } Matched pair
TR_4 BFX88 }

Component changes from Fig. 2 for Fig. 3

R_1	82kΩ	C_5 64μF 10V C426AR/D64
R_2	omitted	
R_3	100kΩ	
R_8	1kΩ	

CLASS A DESIGNS

The other design approach, class A, is now described.

Push-pull class A amplifier

Fig. 4 shows the circuit of a push-pull class A amplifier using two transistors type BD131; it will deliver 2·5W at 3% total harmonic distortion. The BD131 transistors should be of the 'complementary pair' gain specification, that is, having $h_{FE} = 40$ at $I_c = 2A$ $V_{ce} = 1V$.

Component values for Fig. 4

		Mullard Type
C_1	10μF, 16V	C426/E10
C_2	160μF, 2·5V	C426/A160
C_3	125μF, 4V	C426/B125
C_4	500μF, 2·5V	C426/A500
C_5	125μF, 16V	C426/E125
		Mullard Style
R_1	18kΩ	CR16
R_2	47kΩ	CR16
R_3	270Ω, 1W	—
R_4	470Ω, 1W	—
R_5	1·0Ω	CR25
R_6	2·2Ω	CR25
R_7	8·2Ω, 1W	—
R_8	4·7kΩ	CR16
R_9	1kΩ	CR16

TR_1 BD131
TR_2 BD131

In this case the power output is limited by the maximum current capability of the integrated circuit, thus for a given output device type, P_o depends upon V_{cc}. For example, given that,

$$V_{cc} = 20V \quad 24V \quad 30V$$

then

$$P_o = 2·0W \quad 2·5W \quad 3·0W$$

The circuit operation is as follows: the voltage at pin 1 of the IC is well defined, and is used to set up the quiescent

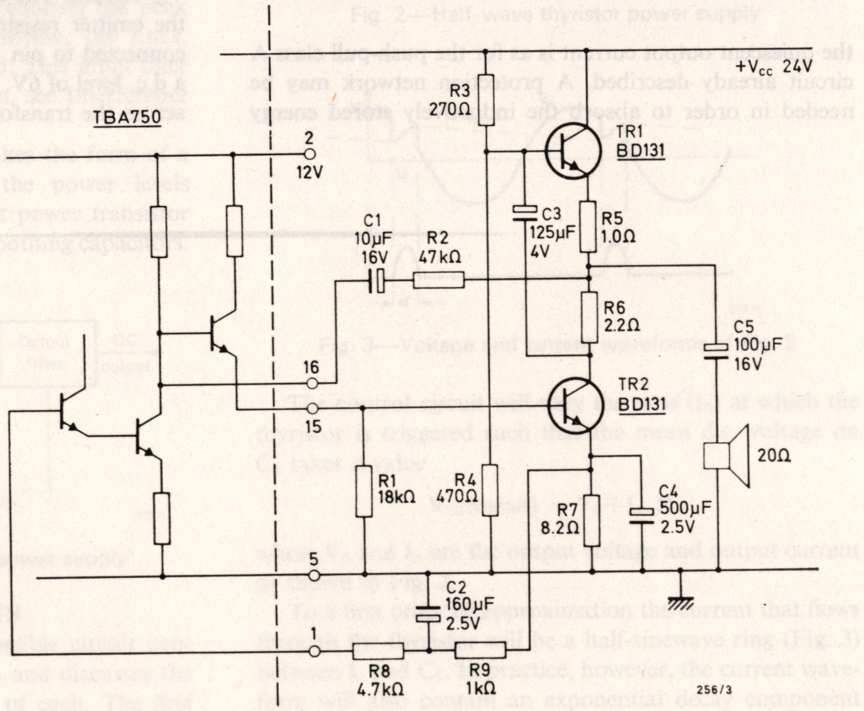

Fig. 4—Class-A push-pull circuit

current in the output pair by defining the voltage across R_7; thus R_7 determines I_q. The value of I_q should be one half of the peak load current.

The output from the integrated circuit is taken from pin 15 to the base of the lower BD131. The drive for the upper BD131 is derived from the collector of the lower BD131.

The resistors R_3 and R_4 set up the mid-point, and R_2 provides a.c. feedback.

Transformer coupled class A circuits

Fig. 5 and 6 show two tentative circuit suggestions for class A output stages with the transformer coupled loads. Fig. 5 is a suggestion for a low V_{cc} circuit. The setting of the quiescent output current is as for the push-pull class A circuit already described. A protection network may be needed in order to absorb the inductively stored energy in the transformer, depending on the ruggedness of the transistor used. Two possible protection arrangements are indicated in Fig. 5 and 6.

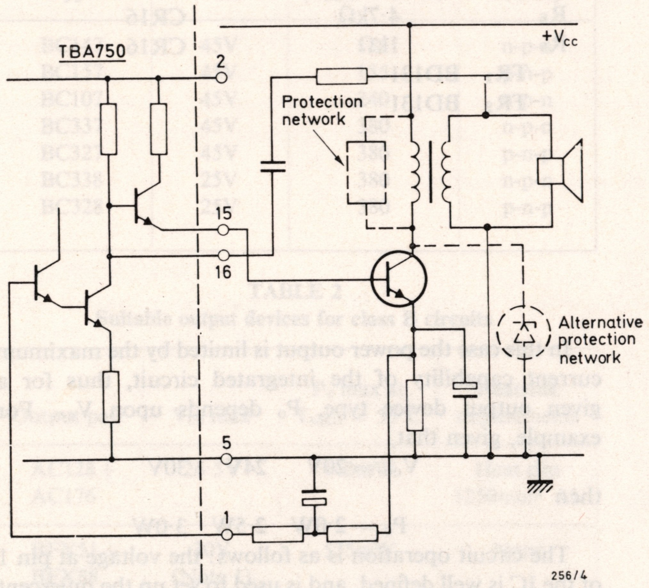

Fig. 5—Low-voltage transformer-coupled class A circuit

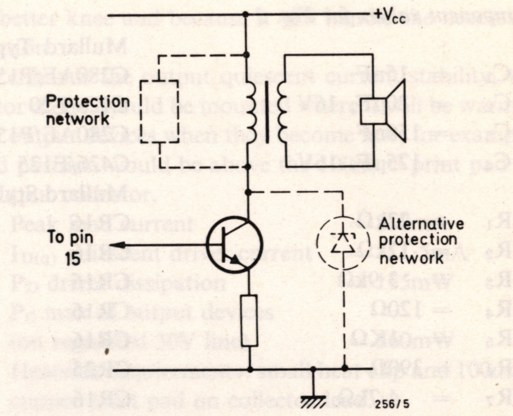

Fig. 6—High-voltage transformer-coupled class A circuit

The one in parallel with the transformer primary could be a VDR or capacitor-resistor network.

The second approach (shown dotted) from the transistor collector to earth is a voltage regulator (zener) diode, chosen to conduct when V_c is greater than about $2\times V_{cc}$.

The strain imposed upon the transistor, and thus the need for protection, is greatly increased if the secondary of the transformer can become open-circuited, as could happen when an extension loudspeaker is employed. When the secondary is open-circuit the whole primary inductance energy has to be absorbed, and the CR network would not be practical as sufficient protection in the open-circuit case would interfere with the normal operation of the circuit.

Fig. 6 shows a circuit which would be suitable for high V_{cc} circuits. The quiescent current is set, in this circuit, by the emitter resistor, the base of the output device being connected to pin 15 of the integrated circuit, maintaining a d.c. level of 6V. Again some form of protection network across the transformer primary would be necessary.

Thyristor power supplies for television receivers: design considerations

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621.311.62[621.382.3]:621.397.13

An economic power supply for television receivers can be constructed by using a thyristor as a controlled rectifier. This article examines various thyristor power supply circuits and discusses practical considerations.

INTRODUCTION

The use of a thyristor as a controlled rectifier enables an h.t. potential in the range 100 to 240V to be derived from the a.c. mains without a transformer and with minimal dissipation in the active devices. This article describes some of the practical circuit configurations and control circuits for thyristor power supplies. Temperature coefficient, interference, isolation for servicing, and safety considerations are also examined.

BASIC CIRCUIT

The basic thyristor power supply circuit is shown in the block diagram of Fig. 1. The control circuit triggers the thyristor in the latter half of the positive half-cycle of the mains waveform so that the mean d.c. voltage presented to the output filter can take a value less than peak mains voltage.

An input filter is used to limit the peak thyristor current to a permitted value, and to reduce r.f. switching currents flowing back into the mains. Also, in reducing the peak thyristor current for a given output current, the r.m.s. current taken from the mains is reduced. This input filter usually takes the form of an LC section, see Interference on page 138.

The output low-pass filter usually takes the form of a single CR combination because of the power levels involved. An active filter using a series power transistor may be used to reduce the size of the smoothing capacitors.

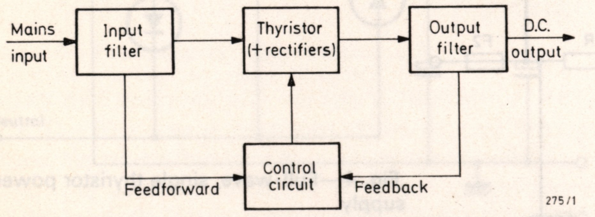

Fig. 1—Block diagram of thyristor power supply

PRACTICAL THYRISTOR CIRCUITS

This section considers some of the possible circuit configurations for thyristor power supplies and discusses the relative advantages and disadvantages of each. The first circuit, Fig. 2, is discussed in somewhat greater detail than the other circuits since the design considerations are common to all.

Simple half-wave thyristor power supply

Consider the circuit of Fig. 2 and the waveforms of Fig. 3. The waveform on capacitor C_1 will be a normal sawtooth, the current only flowing through the thyristor for a few milliseconds between the time when the mains voltage is at peak positive and the time when the mains voltage passes through zero.

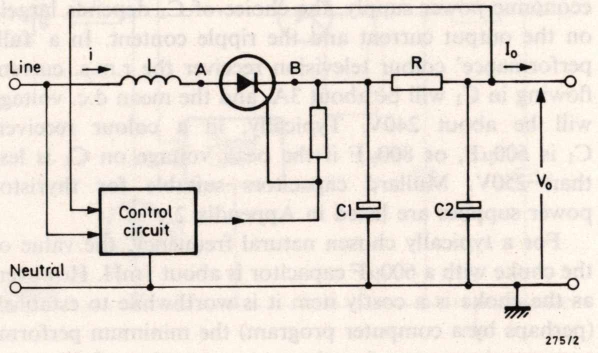

Fig. 2—Half-wave thyristor power supply

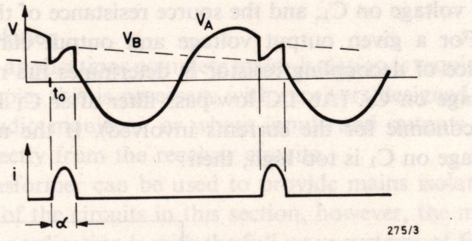

Fig. 3—Voltage and current waveforms of Fig. 2

The control circuit will vary the time (t_o) at which the thyristor is triggered such that the mean d.c. voltage on C_1 takes a value

$$V_{C1}(\text{mean}) = V_o + I_o . R\,,$$

where V_o and I_o are the output voltage and output current as shown in Fig. 2.

To a first order of approximation the current that flows through the thyristor will be a half-sinewave ring (Fig. 3) between L and C_1. In practice, however, the current waveform will also contain an exponential decay component caused by the (unintentional) resistance in series with the choke (partly the series resistance of the choke itself, about 0·5Ω, and partly the source resistance of the mains, about 1Ω). The current waveform will also contain a small 50Hz component. A complete derivation of the thyristor current waveform is given in Appendix 1. Such an expression, when used with similar expressions representing the decay of the circuit when the thyristor is not conducting, can form the basis of a computer program to optimise the values of choke L, capacitors C_1 and C_2, and resistor R. However, a 'trial and error' method of optimising these components can usually result in a satisfactory solution.

The supply authorities recommend that for domestic situations the instantaneous value of current taken from a mains outlet should not exceed the instantaneous value of a 13A r.m.s. sinewave, that is for all values of t

$$i(t) < 13\sqrt{2}\,\sin(100\pi.t).$$

Because of the power levels required in a colour television receiver, this recommendation turns out to be one of the main restrictions on the choice of L and C_1. For a given load current, the peak thyristor current may be reduced by increasing the conduction angle (Fig. 3), and thus a lower natural frequency $(1/2\pi\sqrt{L.C_1})$ is required. For an economic power supply, the choice of C_1 depends largely on the output current and the ripple content. In a 'full-performance' colour television receiver the r.m.s. current flowing in C_1 will be about 3A, and the mean d.c. voltage will be about 240V. Typically, in a colour receiver, C_1 is 600μF, or 800μF if the peak voltage on C_1 is less than 250V. Mullard capacitors suitable for thyristor power supplies are listed in Appendix 2.

For a typically chosen natural frequency, the value of the choke with a 600μF capacitor is about 1mH. However, as the choke is a costly item it is worthwhile to establish (perhaps by a computer program) the minimum performance requirement and optimum construction of this component. The choice of L for a given C_1 depends largely on the output voltage and current, mains voltage, mean d.c. voltage on C_1, and the source resistance of the mains.

For a given output voltage and output current, the choice of decoupling resistor R determines the mean d.c. voltage on C_1. (An LC low-pass filter after C_1 is usually uneconomic for the currents involved). If the mean d.c. voltage on C_1 is too high, then:

1) resistor R has a high dissipation;
2) at low-mains voltage there will be insufficient potential difference to drive current through L, the result is that the output impedance increases and the output voltage finally drops below an acceptable level.

On the other hand, if the mean d.c. voltage on C_1 is too low, then:

1) the thyristor has to be triggered late in the mains cycle, this means that unless unrealistic values of L are used the instantaneous value of the 13A r.m.s. sinewave will be exceeded by the current waveform;
2) a low value of mean d.c. on C_1 is associated with a low value of resistor R, this has an adverse effect on the ripple performance of the power supply.

For television receiver power supplies requiring a d.c. of up to 1A, a mean d.c. voltage of approximately 240V on C_1 at maximum beam current has been found satisfactory with 240V mains. The value of R is then determined from the output voltage and the maximum output current. A practical power supply fulfilling this condition operates satisfactorily from mains voltages in the range 190 to 270V.

The capacitance of C_2 is determined by the permitted amount of ripple voltage on the output of the supply. A thyristor power supply using the circuit of Fig. 2 has two main disadvantages.

1) D.C. is taken from the mains supply; in certain countries this is not permitted.
2) The peak reverse voltage rating, V_{RSM}, of the thyristor has to be rather high.

In practice,

$$\begin{aligned} V_{RSM} = {} & \text{mean d.c. voltage on } C_1 \\ & + \text{peak (negative) mains voltage} \\ & + \text{worst-case negative mains spike} \end{aligned}$$

(Refs. 1 and 2).

However, such a circuit does enable a simple, cheap, and reliable power supply with good transient response to be designed. The circuit also has excellent stabilisation characteristics over a wide range of input voltages and output currents. (Ref. 3.)

Full-wave single thyristor power supply

A full-wave single thyristor power supply circuit configuration is shown in Fig. 4. The use of the diode bridge has several advantages (Ref. 3).

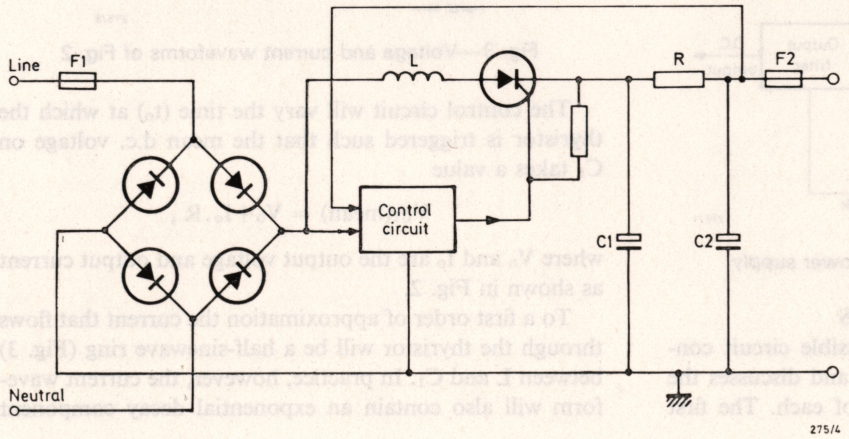

Fig. 4—Full-wave, single thyristor power supply

1) Reduction of reverse blocking requirements of thyristor.
2) No mean d.c. is drawn from the mains supply.
3) Smaller values of C_1 and C_2 may be used for a given output ripple since the rectified waveform is at 100Hz.

There are, however, two major disadvantages (Ref. 4).

1) If for some reason the thyristor fails and becomes short-circuited, peak mains voltage will appear on C_1. In some cases this overvoltage could cause catastrophic failure of the line timebase and video output stages. Even if the fuse F2 blows, the r.m.s. current taken from the mains may be insufficient to blow the mains fuse F1, and peak mains voltage will remain on capacitors C1 and C2. This will be very dangerous unless either C1 and C2 are uneconomically rated on voltage, or some form of comprehensive voltage sensitive cut-out is used.
2) This circuit, regardless of the polarisation of the mains plug, is essentially a 'live-chassis' system and hence the use of an isolating transformer is mandatory for all testing and servicing operations where the chassis is exposed.

Full-wave two thyristor power supply

A full-wave two thyristor power supply circuit configuration is shown in Fig. 5. The use of two diodes and two controlled rectifiers (Ref. 3) in a bridge arrangement has the same advantages as a diode bridge followed by a single thyristor.

1) Reduction of reverse blocking requirements of thyristor.
2) A mean d.c. is not drawn from the mains (provided the circuit is accurately balanced).
3) 100Hz operation, reducing requirements for the ripple filter.

In contrast to the previous circuit with a diode bridge and a single thyristor, a short-circuited thyristor in Fig. 5 will blow the mains fuse F1 safely.

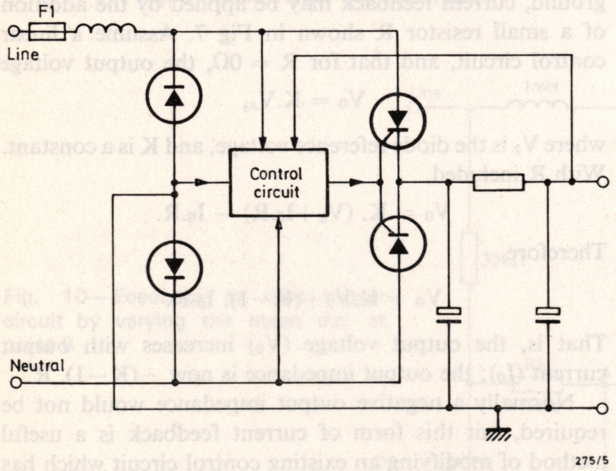

Fig. 5—Full-wave, two thyristor power supply

In practice, the control circuit for such an arrangement has to be very carefully balanced to avoid an appreciable 50Hz component of ripple in the output, and to avoid a mean d.c. being drawn from the mains. This requirement entails either the use of matched pairs of resistors, or two more mains voltage diodes. In either case the cost is increased. Again this is a 'live-chassis' system.

Voltage doubling arrangement

The circuit of Fig. 6 (see also Ref. 4) makes use of a voltage doubling arrangement (capacitor C_1 and d.c. restoring diode D_1) to enable high output voltages to be obtained from a low mains voltage supply (for example, 110V r.m.s.). Also the circuit does not draw d.c. from the mains.

Provided that a suitable control circuit is used, a single switch may be included to disconnect C_1 and D_1 from the circuit. This enables the same circuit to be used on either 110V or 240V mains.

A thermistor or a current limiting resistor must be included in series with C_1 to limit the surge current at switch-on. In any case the capacitor must cope with rather high voltages and ripple currents, and also the forward blocking requirements of the thyristor are somewhat increased.

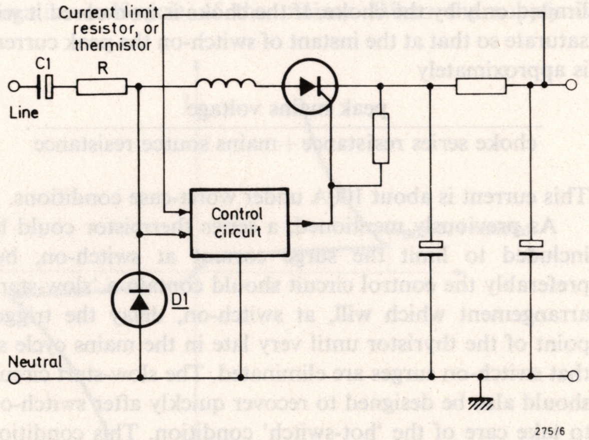

Fig. 6—Voltage doubling arrangement

Mains isolation

In certain applications complete mains isolation is required. For example, this is necessary with receivers designed for use as studio monitors, or where inputs and outputs are taken directly from the receiver circuits.

A transformer can be used to provide mains isolation with any of the circuits in this section, however, the most economic application is with the full-wave systems of Figs. 4 and 5. This is because of the reduced peak and r.m.s. currents in a full-wave system for a given power output. Two factors contribute to a reduction in transformer size and consequently to a more economic system.

1) If the transformer is designed to have an appreciable leakage inductance, a choke is not required.
2) If an electronic slow-start system (see overleaf in the section on control circuits) is not being used, the source resistance of the transformer can be used to limit the current surges at switch-on.

A discussion of the occasional use of a transformer (for servicing purposes) with thyristor supplies is presented on page 139.

CONTROL CIRCUITS

This section considers some of the possible circuits that can be used to trigger the thyristor at a suitable phase angle in the mains cycle, and the various ways by which feedback may be applied. Some practical circuits using the ideas discussed are shown as examples.

For simplicity, the control circuits described apply to Fig. 2, but are basically applicable to all the preceding circuit configurations.

Basic requirements of control circuit

The control circuit should be capable of supplying a short-duration positive going pulse sufficient to trigger the thyristor shortly after the positive peak of the mains supply waveform. Feedback from the output voltage, and usually feedforward from the mains input, vary the phase of the trigger pulse to maintain a constant output voltage for a range of output currents and mains supply potentials.

When the receiver is first switched on, the output voltage will be zero. The control circuit will trigger the thyristor very early in the mains cycle in an effort to achieve the correct output voltage. Thus very large currents can flow, limited only by the choke. If the choke is iron-cored it will saturate so that at the instant of switch-on the peak current is approximately

$$\frac{\text{peak mains voltage}}{\text{choke series resistance}+\text{mains source resistance}}.$$

This current is about 100A under worst-case conditions.

As previously mentioned, a series thermistor could be included to limit the surge current at switch-on, but preferably the control circuit should contain a 'slow-start' arrangement which will, at switch-on, delay the trigger point of the thyristor until very late in the mains cycle so that switch-on surges are eliminated. The slow-start circuit should also be designed to recover quickly after switch-off to take care of the 'hot-switch' condition. This condition exists when the receiver is switched off and then immediately back on again. A thermistor obviously offers no protection against hot-switches.

It is difficult to define thyristor power supplies in terms of unique loop-gain, frequency response and so on. But broadly speaking, the loop gain is expected to be falling at the rate of about 18dB/octave in the region approaching 50Hz because of three main factors which reduce the gain at 'high' frequencies.

1) The 50Hz excitation of the circuit.
2) The time-constant in the ripple filter.
3) The time-constant associated with the timing portion of the trigger circuit.

Thus the gain in the region of (very roughly) 5Hz has to be limited to prevent the circuit from oscillating at this frequency, or from having an oscillatory transient response. In practice it is unlikely that a thyristor power supply will contain components having a frequency-dependent effect below about 5Hz. The net result is that the d.c. loop-gain of a thyristor power supply has a direct effect on the transient response of the circuit. In practice this limits the d.c. loop-gain to a value in the region 20 to 30dB.

Insufficient d.c. loop gain immediately results in insufficient control of the output voltage against variations in mains voltage and output current. So, ideally, a control circuit should incorporate some degree of mains voltage feedforward (outside the control loop), and current feedback as well as voltage feedback. (Non-linear feedback, or feedforward, or both, can sometimes be useful). The overall effect of applying mains voltage feedforward and current feedback is that the d.c. loop-gain may be decreased for a given degree of control against variations in mains voltage and output current. This immediately improves the transient response.

One disadvantage of applying too much mains voltage feedforward, besides giving a decreasing output voltage for increasing mains voltage, is that 'mains-bounce' is amplified. This is a very low-frequency phase modulation of the mains supply which is usually observed in the evening when factories shut down. Too much feedforward results in the mains-bounce disturbance appearing at the output.

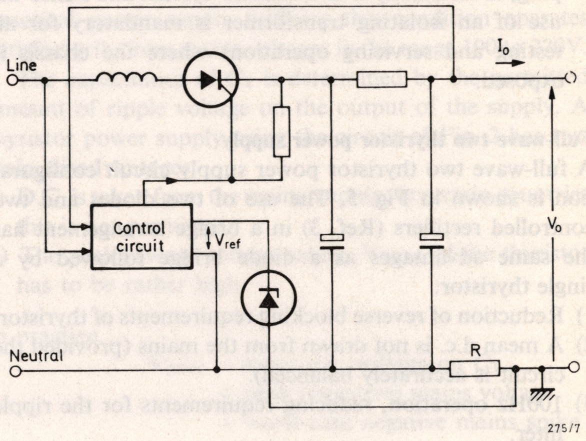

Fig. 7—Current feedback to power supply

Most control circuits have a voltage reference point. If, for example, this is a voltage reference diode connected to ground, current feedback may be applied by the addition of a small resistor R shown in Fig 7. Assume a linear control circuit, and that for R = 0Ω, the output voltage

$$V_0 = K.V_z,$$

where V_z is the diode reference voltage, and K is a constant. With R included,

$$V_0 = K.\,(V_z+I_0.R) - I_0.R.$$

Therefore,

$$V_0 = K.V_z+(K-1).\,I_0.R.$$

That is, the output voltage (V_0) increases with output current (I_0); the output impedance is now $-(K-1)$. R.

Normally a negative output impedance would not be required, but this form of current feedback is a useful method of modifying an existing control circuit which has a high positive output impedance.

Finally, consideration should be given to the temperature coefficient of the control circuit because of the fairly wide range of temperatures encountered within a television receiver cabinet. Temperature coefficients are discussed on page 136.

Diac trigger circuit

Fig. 8 shows a simple unstabilised trigger circuit using a diac BR100. The diac is a bidirectional device which has a negative-resistance region starting at a critical potential called the breakover voltage, V_{BO} (see Fig. 9).

The R_1C_1 network provides a delayed and attenuated version of the mains waveform. Once the potential at point A reaches the breakover voltage of the diac, it is triggered and C_1 partly discharges through the diac. The voltage pulse appearing across R_2 is transmitted through C_2 to the thyristor.

Fig. 10 shows a way of stabilising the circuit of Fig. 8 by varying the mean d.c. at point B. Voltage feedback is applied to point B from the output, and feedforward is supplied from the mains. The voltage reference diode D_1 provides a voltage pedestal for the C_1R_1 network and prevents the diac from triggering in the unwanted direction during the negative portion of the mains cycle. Typical component values are shown. This circuit makes no provision for slow-starting, has a relatively large output impedance, and a poor temperature performance.

Alternatively, the charging rate of capacitor C_1 can be varied by connecting a transistor across it. In practice a reference diode in series with the base-emitter junction is required to swamp V_{BE} temperature coefficient effects. Feedback and feedforward are then applied to the base of the transistor. Fig. 11 shows a typical circuit using a transistor across C_1 for stabilisation (Ref. 3).

In addition to the obvious voltage controlling paths there is an inbuilt current feedback mechanism which operates as follows. When the thyristor is not conducting, the voltage on the cathode of the thyristor decays at a rate dependent on the output current. Thus, a differentiated sawtooth waveform is transmitted through C_2 to point B so that the voltage across R_2 just before the diac triggers is slightly negative, depending on the output current. Hence, the phase angle of triggering is influenced by the load current. The value of R_2 determines the amplitude of the differentiated sawtooth waveform developed and so affects the output impedance. Resistor R_2 also provides damping to point B when there is a transient in the output current.

Effective slow-start may be applied to this circuit by placing an electrolytic capacitor, about 100µF, across the reference diode (and a diode from ground to collector to prevent this capacitor from being discharged during the negative half-cycle of the mains). When the supply is

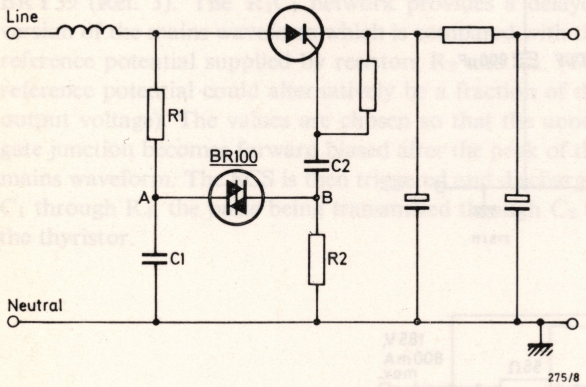

Fig. 8—Diac trigger control circuit

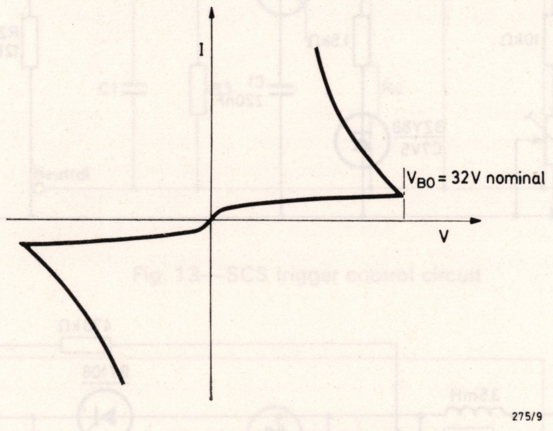

Fig. 9—Diac characteristic

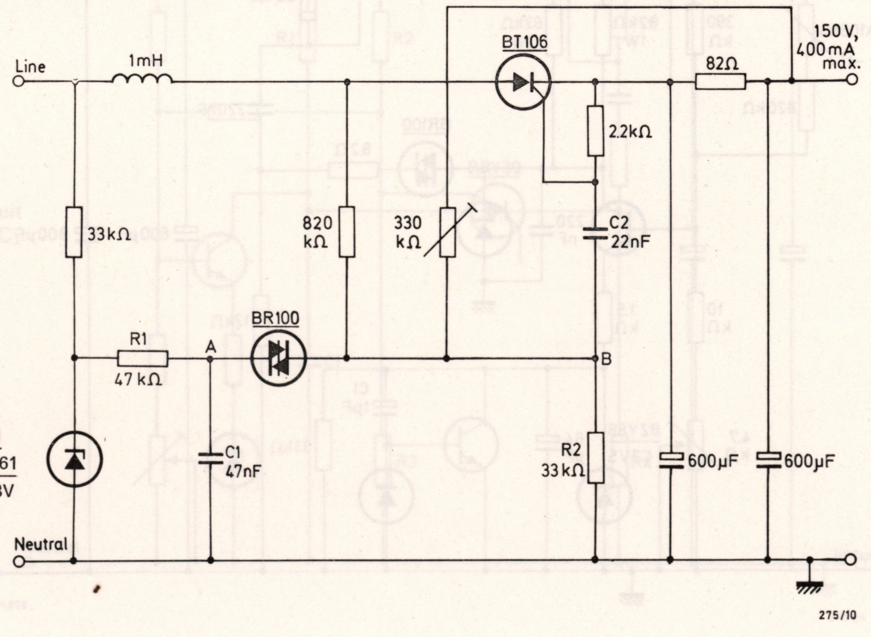

Fig. 10—Feedback to diac trigger circuit by varying the mean d.c. at point B

switched on, the transistor conducts heavily until this capacitor is charged to the reference potential, effectively delaying the trigger point of the thyristor. A resistance of about 22kΩ is required to discharge the capacitor after switch-off. Ideally, the discharge time achieved by this slow-start arrangement, should be small compared with the time taken for the input reservoir capacitor to discharge. By this means, the peak current at switch-on will not greatly exceed the normal operating value. In practice, however, the attainable charge-to-discharge time ratio is limited because the control transistor is required to operate down to relatively low current levels at low mains inputs. Thus, with a low-value discharge resistance, the reference potential would never be reached.

An active circuit for achieving a high charge-to-discharge time ratio has been described (Ref. 5) and a suitable version of this circuit has been incorporated into a thyristor power supply as shown in Fig. 12.

When the receiver is switched on, capacitor C_1 appears as $h_{FE} \times 1\mu F$, thus giving effective slow-start. At switch-off, C_1 discharges rapidly through the 33kΩ resistor and the diode, giving rapid recovery, and therefore protects the thyristor against the hot-switch condition. This circuit has excellent stabilisation and transient response characteristics over a wide range of input voltages and output currents. A typical set of measurements is given in Table 1.

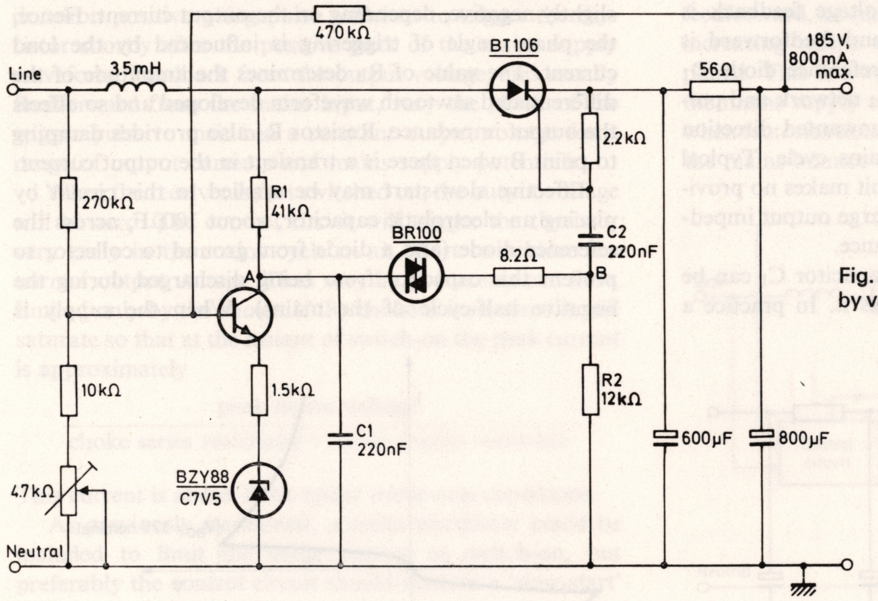
470kΩ
BT106
185V,
800mA
max.
Line
3.5mH
56Ω
2.2kΩ
270kΩ
R1
41kΩ
BR100
C2
220nF
8.2Ω
B
A
10kΩ
1.5kΩ
R2
12kΩ
C1
220nF
600μF
800μF
4.7kΩ
BZY88
C7V5
Neutral
275/11

Fig. 11—Feedback to diac trigger circuit by varying the charging rate of capacitor C_1

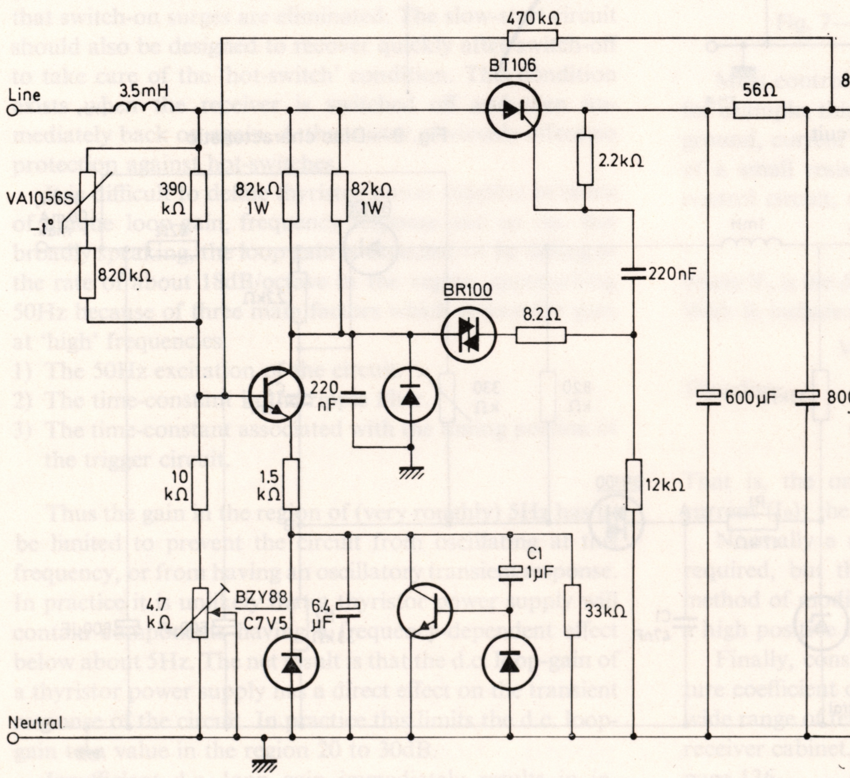
470kΩ
BT106
185V,
800mA
max.
Line
3.5mH
56Ω
2.2kΩ
VA1056S
-t°
390
kΩ
82kΩ
1W
82kΩ
1W
820kΩ
220nF
BR100
8.2Ω
220
nF
600μF
800μF
10
kΩ
1.5
kΩ
12kΩ
C1
1μF
4.7
kΩ
BZY88
C7V5
6.4
μF
33kΩ
Neutral
275/12

Fig. 12—Active slow-start applied to diac trigger circuit

TABLE 1

Measurements on power supply of Fig. 12.

Mains voltage	Load current at 240V mains			
	400mA	600mA	800mA	1000mA
180V	183V	178V	175V	166V
190	184	182	178	174
200	185·5	184	182	178·5
220	186	185	185	184
240	186	185	185	185
260	185·5	185	185	185
270	185·5	185	185	185

Silicon controlled switch (SCS) trigger circuit

Fig. 13 shows a simple trigger circuit using an SCS type BRY39 (Ref. 3). The R_1C_1 network provides a delayed version of the mains waveform which is compared with the reference potential supplied by resistors R_2 and R_3. (The reference potential could alternatively be a fraction of the output voltage). The values are chosen so that the anode gate junction becomes forward biased after the peak of the mains waveform. The SCS is then triggered and discharges C_1 through R_4, the pulse being transmitted through C_2 to the thyristor.

Generally speaking, trigger circuits using an SCS, can be used either on half-wave or full-wave power supplies because C_1 is fully discharged through R_4 every time the SCS is triggered (Ref. 4). A similar trigger circuit using a diac with a full-wave power supply would not operate because of the mean d.c. which would build up on C_1. As with the diac trigger circuit (Fig. 8), the circuit of Fig. 13 may be stabilised by varying the reference potential, or by varying the charging rate of C_1 as a function of the output voltage and mains input voltage, or by varying both.

Fig. 14 shows an arrangement which may be used to control the charging rate of C_1. The control circuit is similar to that of Fig. 11, and slow-start can be applied as in Fig. 12.

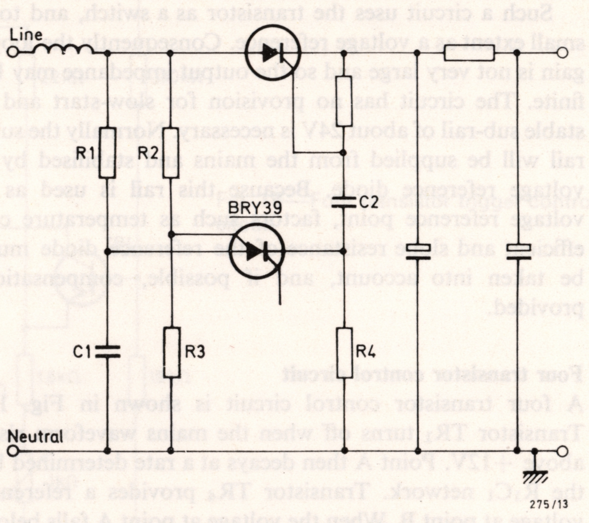
Line
R1
R2
C2
BRY39
C1
R3
R4
Neutral
275/13

Fig. 13—SCS trigger control circuit

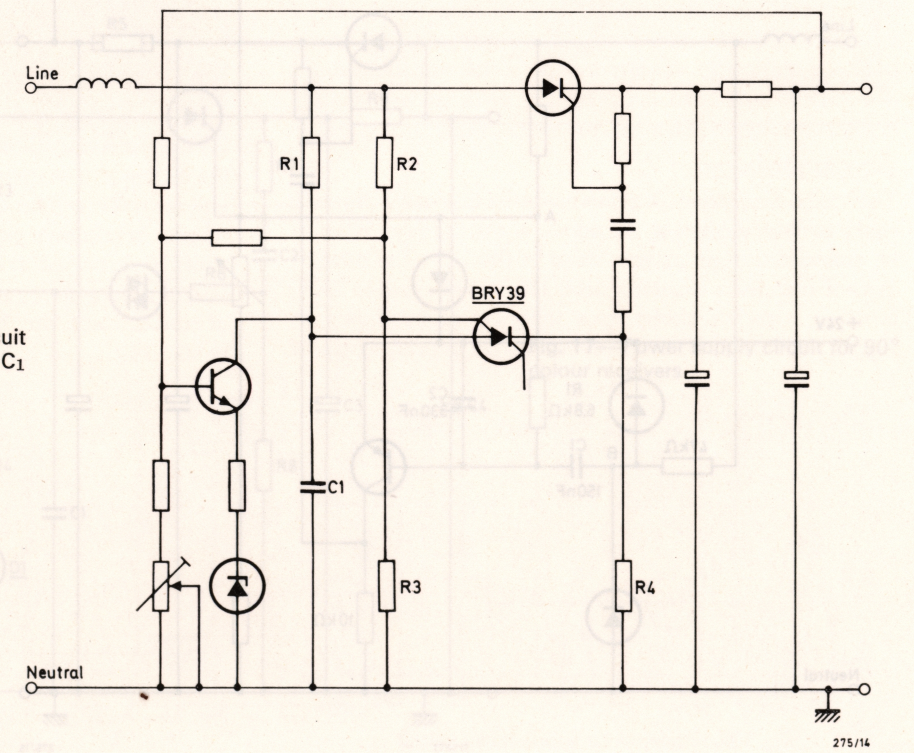
Line
R1
R2
BRY39
C1
R3
R4
Neutral
275/14

Fig. 14—Feedback to SCS trigger circuit by varying the charging rate of capacitor C_1

Single transistor trigger circuit

A suitable trigger pulse may be obtained by differentiating the collector waveform of an overdriven p-n-p transistor. In the circuit of Fig. 15, point B is clamped to zero volts during the negative mains half-cycle. Point A is at about 18V (depending on output voltage) and so the transistor is held hard-on through R_1. When the mains waveform rises above the subrail voltage (24V in this case) point B rises sharply causing the base-emitter junction to become reverse biased through C_1 and C_2. The base voltage then falls at a rate determined by the time-constant $(C_1+C_2)R_1$, and the voltage at point A. When the base-emitter junction becomes forward biased again, the collector rises sharply causing the thyristor to conduct. Mains feedforward is applied to point A.

Such a circuit uses the transistor as a switch, and to a small extent as a voltage reference. Consequently the loop-gain is not very large and so the output impedance may be finite. The circuit has no provision for slow-start and a stable sub-rail of about 24V is necessary. Normally the sub-rail will be supplied from the mains and stabilised by a voltage reference diode. Because this rail is used as a voltage reference point, factors such as temperature coefficient and slope resistance of the reference diode must be taken into account, and if possible, compensation provided.

Four transistor control circuit

A four transistor control circuit is shown in Fig. 16. Transistor TR_1 turns off when the mains waveform rises above +12V. Point A then decays at a rate determined by the R_1C_1 network. Transistor TR_4 provides a reference voltage at point B. When the voltage at point A falls below the voltage at point B, TR_2 turns on and triggers the thyristor through TR_3. Capacitor C_2 provides a sufficiently low turnover frequency for stability, but possibly at the expense of optimum transient response.

The comparatively high d.c. gain of transistor TR_4 enables excellent stabilisation characteristics to be achieved. Capacitor C_2 provides a degree of slow-start. However, this circuit exhibits uninhibited hot-switch behaviour and requires a stable +12V supply rail. Therefore, as in Fig. 15, the sub-rail (in this case +12V) is again used as a reference potential and any temperature coefficient effects must be taken into account.

TEMPERATURE COEFFICIENT

In colour television receivers, the power supply rail to a transistorised line timebase may have to be stabilised to within about 2 or 3 per cent. of its nominal value. Usually the output voltage can be set up by a suitable potentiometer, but temperature coefficient effects must be minimised for subsequent operation of the receiver. Using the circuit of Fig. 17 as an example, the output voltage is mainly a function of resistor values (and ratios), but the temperature coefficients of the reference diode, the diac, and the V_{BE} of the control transistor also play an important part. The choice of reference diode gives some degree of control over the temperature coefficient of the overall circuit. Assuming zero coefficient resistors, a suitably chosen reference diode can compensate for the temperature coefficient of the diac and the V_{BE} of the control transistor.

The circuit in Fig. 17 is of a simple power supply used for 90° colour receivers. The temperature coefficient of each resistor can be determined by changing each resistor value in turn by 10% and making note of the subsequent effect on the magnitude and direction of the change in output voltage. A typical set of measurements is summarised in Table 2.

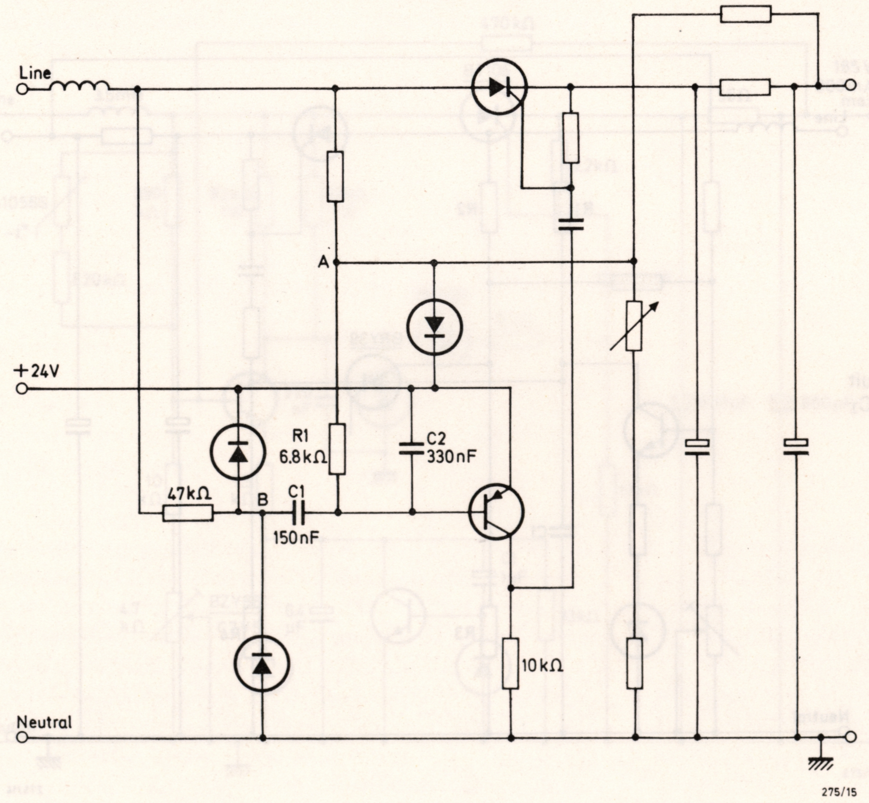

Fig. 15—Single transistor trigger control circuit

The other components in the thyristor power supply do not have a significant effect on the output voltage. The V_{BE} of the control transistor is included in combination with the reference voltage of D_1 in Table 2.

If all the resistors in this circuit had the same temperature coefficient, their effects with a rise in ambient temperature would tend to balance out. However, reference to the published data shows that metal-oxide and metal-film resistors have a quoted spread in temperature coefficient of ±250ppm/degC. The worst combination of coefficients leads to a ±20V change in output voltage for a 50degC rise in ambient temperature.

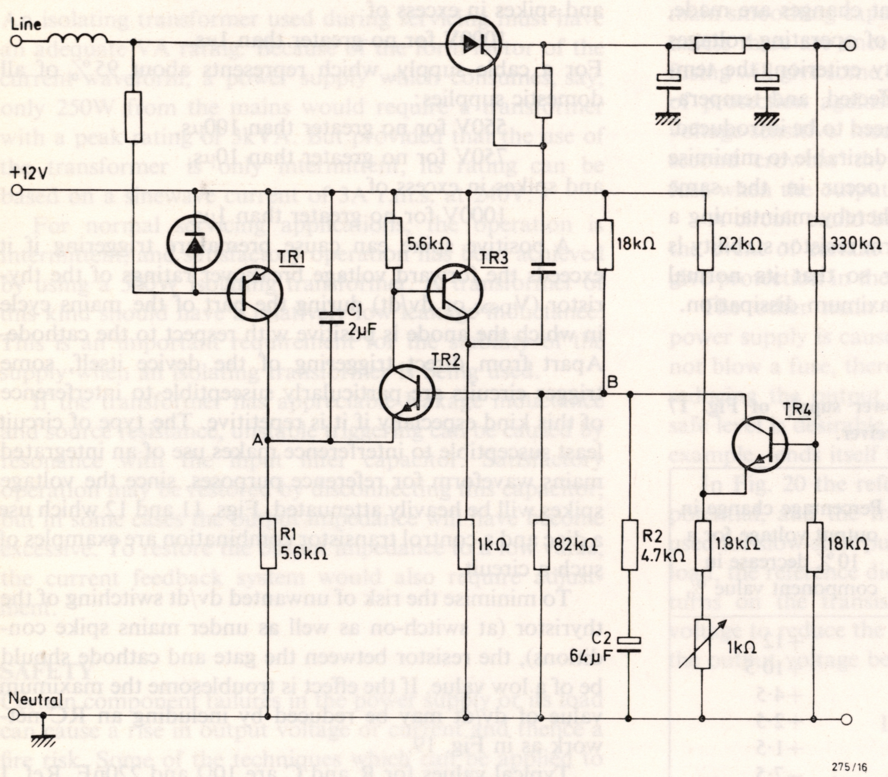

Fig. 16—Four transistor trigger control circuit

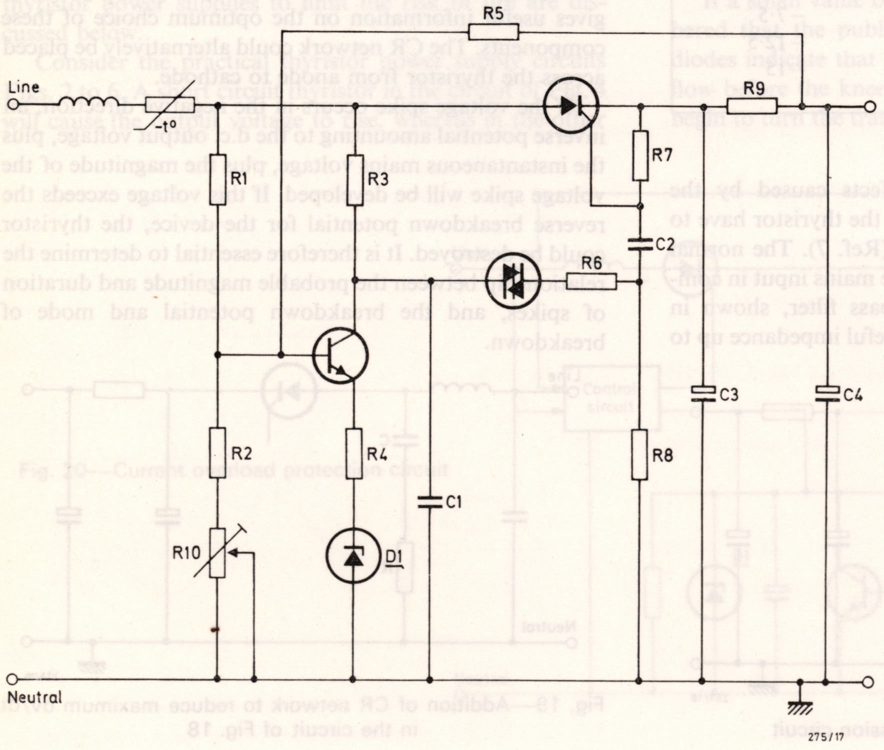

Fig. 17—Power supply circuit for 90° colour receivers

Solid-carbon and carbon-film resistors both have a predictable temperature coefficient (typically −260 ppm/degC, depending on resistance value), but only carbon-film resistors have adequate long term stability (Ref. 6). Taking the published data for the Mullard CR25 and CR68 carbon-film resistors, the calculated overall temperature coefficient of the power supply in Fig. 17 is about +1V for a 50 degC rise in ambient temperature.

These results apply for the particular choice of component values in Table 2. If significant changes are made, perhaps to achieve a different range of operating voltages or currents, or to a different stability criterion, the temperature drift cancellation will be affected, and temperature compensating components may need to be introduced.

The use of carbon-film resistors is desirable to minimise long-term drift effects which will occur in the same direction and at a predictable rate, thereby maintaining a degree of tracking. Optimum long-term resistor stability is achieved by operating each resistor so that its normal dissipation is well below its rated maximum dissipation.

TABLE 2

Typical set of measurements for the power supply of Fig. 17 used in a 90° colour receiver.

Components, and nominal value		Percentage change in output voltage for a 10% decrease in component value %
R_2	9·1kΩ	+12
R_3	33kΩ	+10·5
R_{10} 4·7kΩ preset (3·55kΩ in this case)		+4·5
Diac breakover voltage (32V)		+2·5
C_1	220nF	+1·5
R_4	820Ω	−7·5
R_5	390kΩ	−7·5
R_1	390kΩ	−12·5
D_1 reference voltage (7·5V)		−15

INTERFERENCE

The radio-frequency interference effects caused by the sharp switching edges at the anode of the thyristor have to be attenuated to an acceptable level (Ref. 7). The normal method is to use a capacitor across the mains input in combination with the choke as a low-pass filter, shown in Fig. 18. A C-core choke will have a useful impedance up to about 30MHz.

Typically C is a 220nF, 1kV capacitor. To meet extreme radiation requirements a second LC section may be necessary. Large voltage spikes caused by a switching transient elsewhere in the vicinity can come through the mains supply and cause interference (Ref. 2). For a domestic overhead supply the worst-case spikes can be summarised as being:

750V for no greater than 100μs,
950V for no greater than 10μs,

and spikes in excess of

1000V for no greater than 1μs.

For a cable supply, which represents about 95% of all domestic supplies:

550V for no greater than 100μs,
750V for no greater than 10μs,

and spikes in excess of

1000V for no greater than 1μs.

A positive spike can cause premature triggering if it exceeds the forward voltage breakover ratings of the thyristor (V_{DSM} or dv/dt) during the part of the mains cycle in which the anode is positive with respect to the cathode. Apart from direct triggering of the device itself, some trigger circuits are particularly susceptible to interference of this kind especially if it is repetitive. The type of circuit least susceptible to interference makes use of an integrated mains waveform for reference purposes, since the voltage spikes will be heavily attenuated. Figs. 11 and 12 which use a diac and a control transistor combination are examples of such a circuit.

To minimise the risk of unwanted dv/dt switching of the thyristor (at switch-on as well as under mains spike conditions), the resistor between the gate and cathode should be of a low value. If the effect is troublesome the maximum value of dv/dt may be reduced by including an RC network as in Fig. 19.

Typical values for R and C are 10Ω and 220nF. Ref. 1 gives useful information on the optimum choice of these components. The CR network could alternatively be placed across the thyristor from anode to cathode.

If the voltage spike occurs in the negative direction, an inverse potential amounting to the d.c. output voltage, plus the instantaneous mains voltage, plus the magnitude of the voltage spike will be developed. If this voltage exceeds the reverse breakdown potential for the device, the thyristor could be destroyed. It is therefore essential to determine the relationship between the probable magnitude and duration of spikes, and the breakdown potential and mode of breakdown.

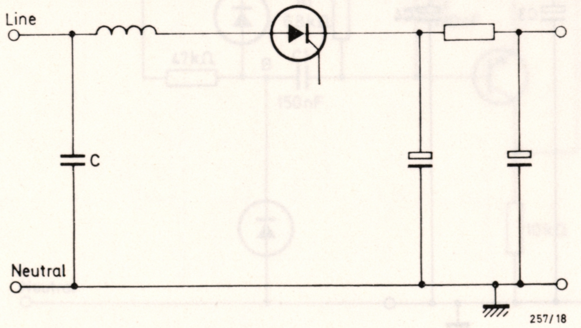

Fig. 18—Interference suppression circuit

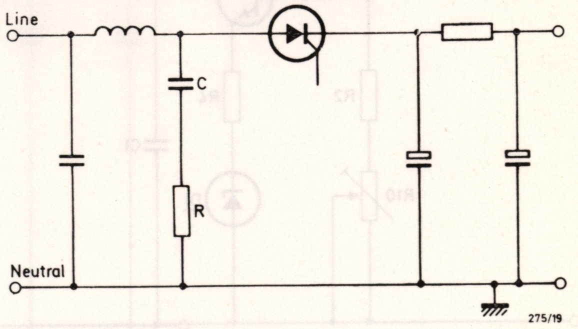

Fig. 19—Addition of CR network to reduce maximum dv/dt in the circuit of Fig. 18

If there is a very large di/dt through the thyristor once it has been triggered (either normally or by a positive spike), and before a uniform current density has been established across the junction, there may be local heating effects at the point where the current flow starts. Exceeding the permitted di/dt may cause a thyristor to fail. Correct specification of the choke will guard against such a possibility.

ISOLATION FOR SERVICING

An isolating transformer used during servicing must have an adequate VA rating. Because of the form-factor of the current waveform, a power supply which consumes, say, only 250W from the mains would require a transformer with a peak rating of 3kVA. But provided that the use of the transformer is only intermittent, its rating can be based on a sinewave current of 3A r.m.s. at 240V.

For normal servicing applications, the operation is intermittent, and satisfactory operation has been achieved by using a 500W isolating transformer. A transformer of this kind should have a relatively low leakage inductance. This is an important requirement for the stability of the supply when an isolating transformer is being used.

If the transformer has appreciable leakage inductance and source resistance, unstable triggering can be caused by resonance with the input filter capacitor. Satisfactory operation may be restored by disconnecting this capacitor, but in some cases the output impedance will have become excessive. To restore the output impedance to a low value, the current feedback system would also require adjustment.

SAFETY

Certain component failures in the power supply or its load can cause a rise in output voltage or current and thence a fire risk. Some of the techniques which can be applied to thyristor power supplies to limit the risk of fire are discussed below.

Consider the practical thyristor power supply circuits Figs. 2 to 6. A short circuit thyristor in the circuit of Fig. 4 will cause the output voltage to rise, whereas in the other circuits the mains fuse will blow provided it is suitably rated. Other conditions which may cause the output voltage to rise are a failed control circuit or removal of the load from the output of a power supply having insufficient feedback to hold the output voltage to a safe level. Overvoltage may be produced at the output of the power supply, or at the input filter capacitor, or at both, depending on the location of the fault.

In the event of a prolonged overvoltage condition, the main smoothing capacitors in the power supply are at risk unless these are uneconomically rated on voltage. Careful fusing will give some degree of protection. Another method of protection against overvoltage involves the use of a voltage-sensitive mechanical cut-out, or alternatively, a second 'crowbar' thyristor can be used to blow the mains fuse when the output rises above a safe level.

A circuit could be used to disable the trigger circuit in the event of certain fault conditions, but this would not give protection in the case of a failed thyristor.

The other main class of malfunction of a thyristor power supply is caused by a current overload. If this does not blow a fuse, there may be a fire risk. Some method of reducing the output voltage when the current exceeds a safe level is desirable. The slow-start circuit of Fig. 12, for example, lends itself to this application.

In Fig. 20 the reference diode D_1 provides a reference potential, and the transistor and associated circuitry are used for slow-start purposes. In the event of a current overload, the reference diode D_2 comes into conduction, which turns on the transistor and thus reduces the reference voltage to reduce the output voltage. The current at which the output voltage begins to fall, is given by the equation,

$$I = \frac{V_{Z(D2)} + V_{BE}}{R}.$$

If a small value of R is necessary, it should be remembered that the published data for low-voltage reference diodes indicate that a certain amount of current starts to flow before the knee voltage is reached. This current will begin to turn the transistor on before a current overload is

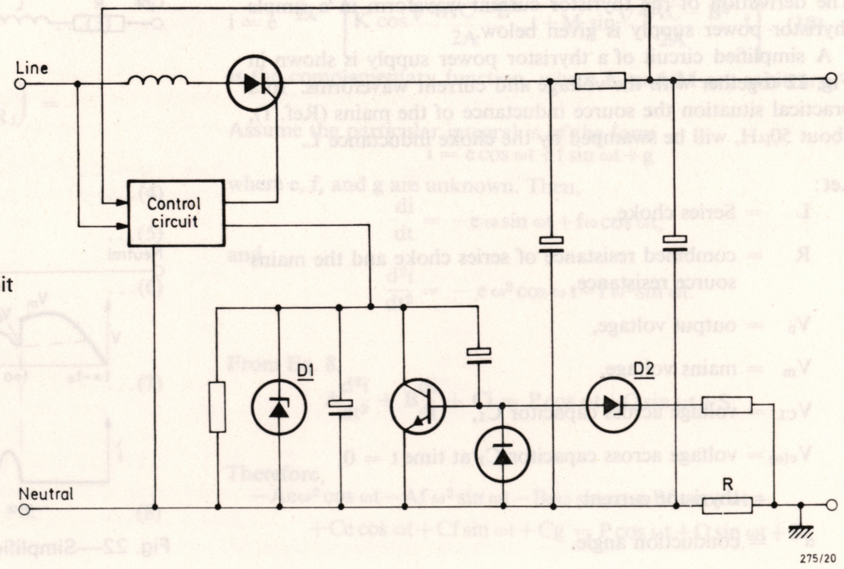

Fig. 20—Current overload protection circuit

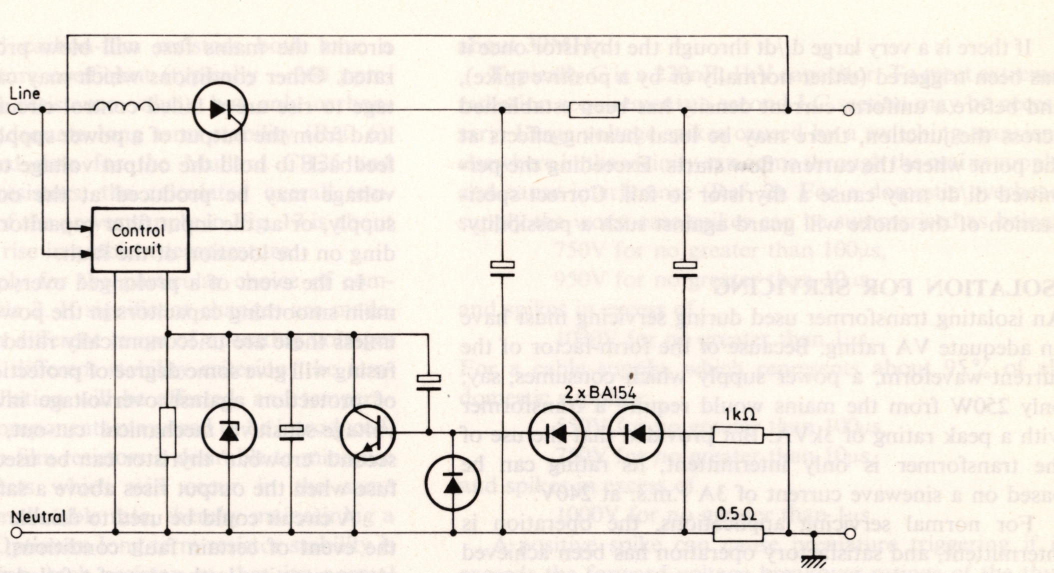

Fig. 21—Current overload protection circuit; reduces output voltage of the circuit in Fig. 12 by 300V/A

actually reached, resulting in distorted output impedance characteristics. This can be avoided by using a forward-biased diode instead of a reference diode, as in Fig. 21.

This arrangement reduced the output voltage of the circuit of Fig. 12 at the rate of 300V/A starting from 1·2A output current.

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APPENDIX 1

Derivation of thyristor current waveform equation

The derivation of the thyristor current waveform in a simple thyristor power supply is given below.

A simplified circuit of a thyristor power supply is shown in Fig. 22 together with the voltage and current waveforms. In a practical situation the source inductance of the mains (Ref. 1), about 50μH, will be swamped by the choke inductance L.

Let:

L = Series choke,

R = combined resistance of series choke and the mains source resistance,

V_o = output voltage,

V_m = mains voltage,

V_{C1} = voltage across capacitor C_1,

$V_{c(o)}$ = voltage across capacitor C_1 at time $t = 0$,

= thyristor current,

α = conduction angle.

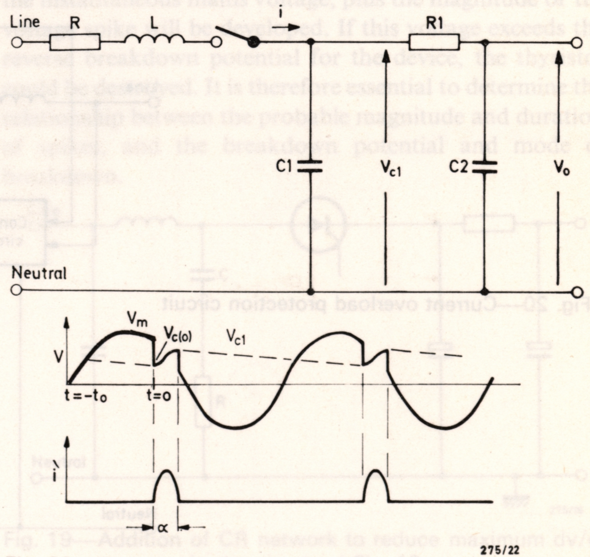


Fig. 22—Simplified power supply circuit, with waveforms

For convenience, assume that the switch is closed at $t = 0$. Therefore,

$$V_m = \hat{V}_m.\sin\omega(t+t_o),$$

and

$$V_m - V_{C1} = iR + L\frac{di}{dt} \qquad \text{(see Fig. 22).}$$

Differentiating,

$$L\frac{d^2i}{dt^2} + R\frac{di}{dt} = \frac{dV_m}{dt} - \frac{dV_{C1}}{dt}.$$

But,

$$\frac{dV_{C1}}{dt} = \frac{1}{C_1}\left(i - \frac{V_{C1}-V_0}{R_1}\right).$$

Therefore,

$$L\frac{d^2i}{dt^2} + R\frac{di}{dt} = \frac{dV_m}{dt} - \frac{1}{C_1}\left(i - \frac{V_{C1}-V_0}{R_1}\right).$$

This assumes that V_o remains substantially constant while the thyristor is conducting, and in practice this is correct to within about 0·1%.

Therefore,

$$L\frac{d^2i}{dt^2} + R\frac{di}{dt} = \hat{V}_m.\omega\cos\omega(t+t_o) -$$

$$-\frac{i}{C_1} - \frac{V_0}{C_1R_1} + \frac{1}{C_1R_1}\left(V_m - L\frac{di}{dt} - iR\right).$$

Hence,

$$L\frac{d^2i}{dt^2} + \frac{di}{dt}\left(R + \frac{L}{C_1R_1}\right) + i\left(\frac{1}{C_1} + \frac{R}{C_1R_1}\right) =$$

$$= \hat{V}_m.\omega\cos\omega(t+t_o) + \frac{\hat{V}_m}{C_1R_1}.\sin\omega(t+t_o) - \frac{V_0}{C_1R_1}.$$

Consider the r.h.s,

$$\hat{V}_m.\omega\cos\omega(t+t_o) + \frac{\hat{V}_m}{C_1R_1}.\sin\omega(t+t_o) - \frac{V_0}{C_1R_1} =$$

$$= \hat{V}_m.\omega\cos\omega t_o.\cos\omega t + \frac{\hat{V}_m}{C_1R_1}.\sin\omega t_o.\cos\omega t -$$

$$-\hat{V}_m.\omega\sin\omega t_o.\sin\omega t + \frac{\hat{V}_m}{C_1R_1}.\cos\omega t_o.\sin\omega t - \frac{V_0}{C_1R_1}.$$

Let

$$P = \hat{V}_m\left(\omega\cos\omega t_o + \frac{1}{C_1R_1}.\sin\omega t_o\right), \qquad \ldots(1)$$

$$Q = \hat{V}_m\left(-\omega\sin\omega t_o + \frac{1}{C_1R_1}.\cos\omega t_o\right), \qquad \ldots(2)$$

and

$$S = -\frac{V_0}{C_1R_1}. \qquad \ldots(3)$$

Then

$$L\frac{d^2i}{dt^2} + \frac{di}{dt}\left(R + \frac{L}{C_1R_1}\right) + i\left(\frac{1}{C_1} + \frac{R}{C_1R_1}\right) =$$

$$= P\cos\omega t + Q\sin\omega t + S.$$

Now, let

$$A = L, \qquad \ldots(4)$$

$$B = R + \frac{L}{C_1R_1}, \qquad \ldots(5)$$

$$C = \frac{1}{C_1} + \frac{R}{C_1R_1}, \qquad \ldots(6)$$

and let

$$\frac{di}{dt}(\text{at } t = 0) = \frac{\hat{V}_m\sin\omega t_o - V_{c(o)}}{L} = N. \qquad \ldots(7)$$

Thus,

$$A\frac{d^2i}{dt^2} + B\frac{di}{dt} + Ci = P\cos\omega t + Q\sin\omega t + S. \qquad \ldots(8)$$

The auxiliary equation is

$$Am^2 + Bm + C = O,$$

which has roots m_1 and m_2

$$= \frac{-B \pm \sqrt{B^2-4AC}}{2A}.$$

Hence the complementary function is

$$i = He^{m_1t} + Je^{m_2t} \qquad \ldots(9)$$

where H and J are arbitrary.

In this case (by inspection),

$$B^2 < 4AC.$$

Therefore,

$$m_1 = -\frac{B}{2A} + j.\frac{\sqrt{4AC-B^2}}{2A},$$

and

$$m_2 = -\frac{B}{2A} - j.\frac{\sqrt{4AC-B^2}}{2A}.$$

Substituting for m_1 and m_2 in Eq. 2,

$$i = He^{-\frac{B}{2A}.t} \times e^{j\frac{\sqrt{4AC-B^2}}{2A}.t} +$$

$$+ Je^{-\frac{B}{2A}.t} \times e^{-j\frac{\sqrt{4AC-B^2}}{2A}.t}.$$

Therefore,

$$i = e^{-\frac{B}{2A}.t}\left[H\cos\frac{\sqrt{4AC-B^2}}{2A}.t + jH\sin\frac{\sqrt{4AC-B^2}}{2A}.t + \right.$$

$$\left. + J\cos\frac{\sqrt{4AC-B^2}}{2A}.t - jJ\sin\frac{\sqrt{4AC-B^2}}{2A}.t\right].$$

Put

$$K = H + J,$$

and

$$M = j(H - J).$$

Then,

$$i = e^{-\frac{B}{2A}.t}\left[K\cos\frac{\sqrt{4AC-B^2}}{2A}.t + M\sin\frac{\sqrt{4AC-B^2}}{2A}.t\right] \ldots(10)$$

is the complementary function, where K and M are arbitrary.

Assume the particular integral is of the form

$$i = e\cos\omega t + f\sin\omega t + g$$

where e, f, and g are unknown. Then,

$$\frac{di}{dt} = -e\omega\sin\omega t + f\omega\cos\omega t,$$

and

$$\frac{d^2i}{dt^2} = -e\omega^2\cos\omega t - f\omega^2\sin\omega t.$$

From Eq. 8,

$$A\frac{d^2i}{dt^2} + B\frac{di}{dt} + Ci = P\cos\omega t + Q\sin\omega t + S.$$

Therefore,

$$-Ae\omega^2\cos\omega t - Af\omega^2\sin\omega t - Be\omega\sin\omega t + Bf\omega\cos\omega t +$$

$$+ Ce\cos\omega t + Cf\sin\omega t + Cg = P\cos\omega t + Q\sin\omega t + S.$$

Equating coefficients,

$$g = S/C \qquad \ldots(11)$$

$$P = Ce + Bf\omega - Ae\omega^2 \qquad \ldots(12)$$

$$Q = Cf - Be\omega - Af\omega^2 \qquad \ldots(13)$$

and from Eq. 12,

$$e = \frac{P - Bf\omega}{C - A\omega^2}. \qquad \ldots(14)$$

Substituting for e in Eq. 13,

$$Q = Cf - B\omega\,\frac{P - Bf\omega}{C - A\omega^2} - Af\omega^2.$$

Therefore,

$$Q(C - A\omega^2) = Cf(C - A\omega^2) - B\omega P + fB^2\omega^2 - fA\omega^2(C - A\omega^2).$$

Hence,

$$f = \frac{Q(C - A\omega^2) + B\omega P}{C(C - A\omega^2) + B^2\omega^2 - A\omega^2(C - A\omega^2)} \qquad \ldots(15)$$

Since e is known in terms of f (Eq. 14), and g in terms of S and C, the particular integral ($i = e\cos\omega t + f\sin\omega t + g$) can be determined in terms of the coefficients of Eq. 8.
Thus the general solution is,

$$i = e^{-\frac{B}{2A}.t}\left[K\cos\frac{\sqrt{4AC - B^2}}{2A}.t + M.\sin\frac{\sqrt{4AC - B^2}}{2A}.t\right] + $$
$$+ e\cos\omega t + f\sin\omega t + g. \qquad \ldots(16)$$

K and M are the only unknowns remaining.

At time $t = 0$, $i = 0$ and $\frac{di}{dt} = N$.

Putting 0 for t in Eq. 16 gives,

$$0 = K + e + g.$$

Therefore,

$$K = -(e + g). \qquad \ldots(17)$$

Differentiating Eq. 16,

$$\frac{di}{dt} = -K.e^{-\frac{B}{2A}.t}.\frac{\sqrt{4AC - B^2}}{2A}.\sin\frac{\sqrt{4AC - B^2}}{2A}.t -$$
$$-K.\frac{B}{2A}.e^{-\frac{B}{2A}.t}.\cos\frac{\sqrt{4AC - B_2}}{2A}.t \; +$$
$$+ M e^{-\frac{B}{2A}.t}.\frac{\sqrt{4AC - B^2}}{2A}.\cos\frac{\sqrt{4AC - B^2}}{2A}.t \; -$$
$$-M.\frac{B}{2A}.e^{-\frac{B}{2A}.t}.\sin\frac{\sqrt{4AC - B^2}}{2A}.t -$$
$$-\omega e\sin\omega t \qquad + \omega f\cos\omega t.$$

Now $\frac{di}{dt} = N$ at $t = 0$,

therefore,

$$N = -K.\frac{B}{2A} + M.\frac{\sqrt{4AC - B^2}}{2A} + \omega f.$$

Thus

$$M = \frac{(N + (B/2A)\,K - \omega f).\;2A}{\sqrt{4AC - B^2}}. \qquad \ldots(18)$$

Therefore the current waveform through the thyristor is,

$$i = e^{-\frac{B}{2A}.t}\left[K\cos\frac{\sqrt{4AC - B^2}}{2A}.t + M\sin\frac{\sqrt{4AC - B^2}}{2A}.t\right] +$$
$$+ e\cos\omega t + f\sin\omega t + g,$$

where ω is 100π and where
A is calculated from Eq. 4,
B from Eq. 5,
C from Eq. 6,
K from Eq. 17,
M from Eq. 18,

e from Eq. 14,
f from Eq. 15,
g from Eq. 11,
P from Eq. 1,
Q from Eq. 2,
N from Eq. 7,
S from Eq. 3.

APPENDIX 2

Special Mullard electrolytic capacitors suitable for use in television power supplies.

Capacitor value (μF)	Voltage and ripple current ratings* (V)	(A)	Type No.
(400+400)	200	1·6	057 90048
1000	200	1·2	060 90022
600	250	3·5	060 90018
800	250	3·5	067 90045
600	300	3·0	060 90017
(200+200+100)	350	1·1	067 90046

*Ripple current is measured at 50Hz for all of these capacitors.

Mullard electrolytic capacitors for general use in power supplies are available in the 071 and 072 series, and the 106 and 107 series (see published data).

APPENDIX 3

Loop gain measurements

The frequency and phase response of the loop-gain of a thyristor power supply may be measured as in Fig. 23.

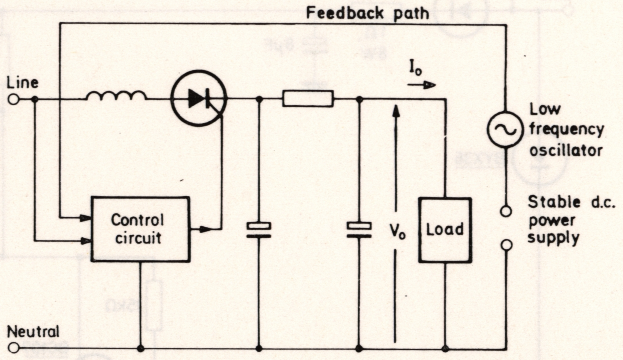


Fig. 23—Circuit for measuring loop-gain

The low-frequency oscillator should be able to provide, say, 1V peak-to-peak from 0·1 to 100Hz. Fig. 24 shows a simple mixer which combines the outputs of two oscillators set at about 1kHz to provide a low-frequency beat signal.

Input A should be overdriven, or supplied with square-waves of sufficient amplitude to switch the current flowing in transistor TR_3 between transistors TR_1 and TR_2. Transistor TR_4 provides some level shifting and gain.

In practice a good sinewave with a period of 15 seconds is easily obtainable using simple audio oscillators.

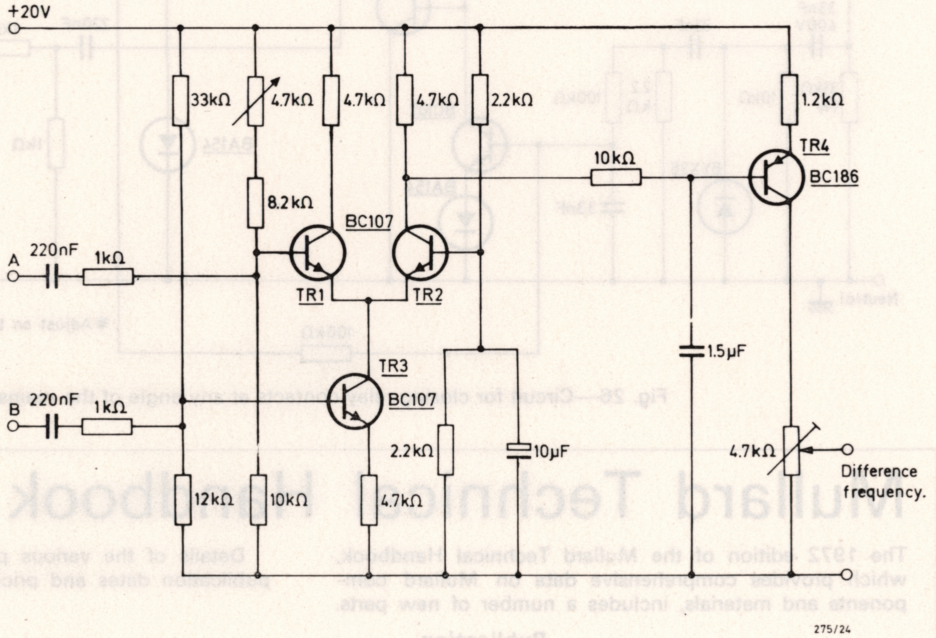


Fig. 24—Simple mixer circuit for producing low-frequency beat signals

APPENDIX 4

Test circuit for dV/dt response

It became necessary to switch on the mains supply at a given phase angle of the mains waveform in order to test for dv/dt effects, overshoot, and current surges at switch-on.

To provide realistic values of dv/dt corresponding to operation of a mains switch, the system must be centred upon a relay. The method used is shown in Fig. 25.

The position of the output pulse from the monostable can be varied to occur at any point in the mains cycle. After the switch is closed (manually), the thyristor is off, and nothing happens. The thyristor is triggered once it receives a pulse from the monostable, and then the relay contacts begin to close. The phase angle at which the relay contacts close is thus purely determined by the monostable setting. The relay contacts remain closed until the switch is opened again. Fig. 26 shows one version of the system.

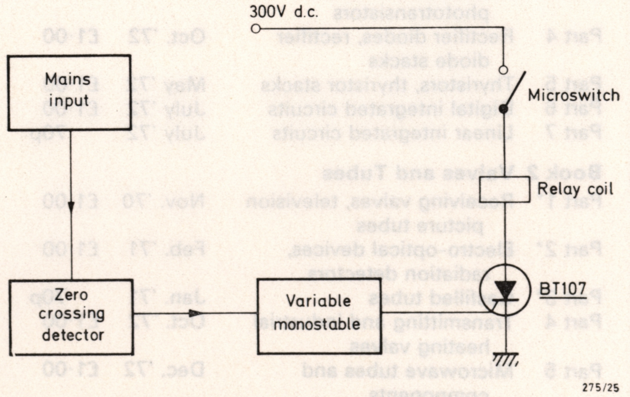
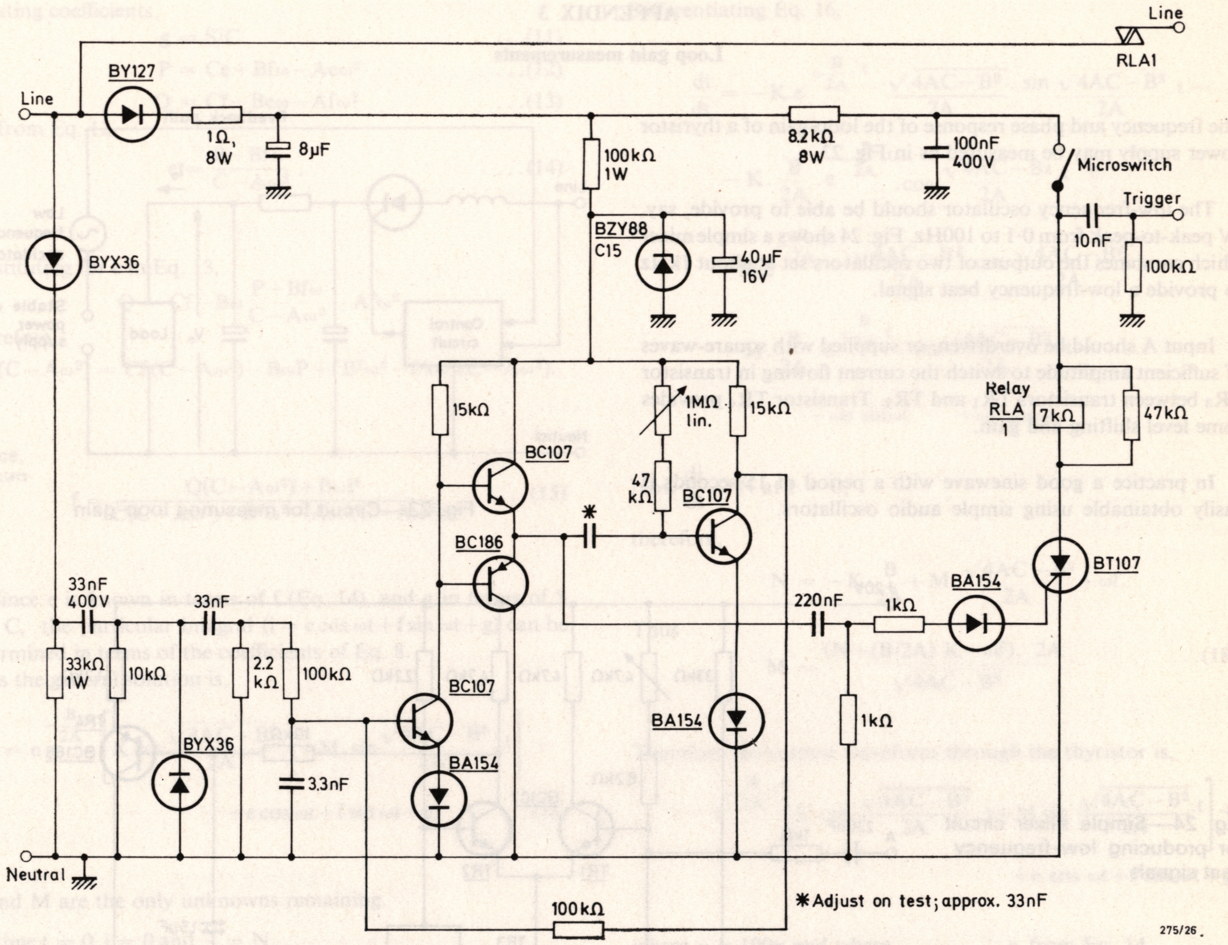


Fig. 25—Schematic diagram for dV/dt test circuit


Line
BY127
1Ω, 8W
8μF
BYX36
100kΩ 1W
BZY88 C15
40μF 16V
8.2kΩ 8W
100nF 400V
Line
RLA1
Microswitch
Trigger
10nF
100kΩ
Relay RLA 1
7kΩ
47kΩ
15kΩ
BC107
1MΩ lin.
15kΩ
47 kΩ
BC107
*
BC186
BT107
220nF
1kΩ
BA154
33nF 400V
33nF
33kΩ 1W
10kΩ
2.2 kΩ
100kΩ
BC107
BYX36
BA154
3.3nF
BA154
1kΩ
Neutral
100kΩ
275/26

*Adjust on test; approx. 33nF

Fig. 26—Circuit for closing relay contacts at any angle of the mains cycle

Mullard Technical Handbook

The 1972 edition of the Mullard Technical Handbook, which provides comprehensive data on Mullard components and materials, includes a number of new parts.

Details of the various parts, together with scheduled publication dates and prices are given below.

		Publication Date	Price
Book 1	**Semiconductor Devices**		
Part 1	Transistors (AC127 to BF**)	Oct. '72	£1·00
Part 2	Transistors (BF** to OC207, 2N types)	Oct. '72	£1·00
Part 3	Diodes, photodiodes, phototransistors	Dec. '72	£1·00
Part 4	Rectifier diodes, rectifier diode stacks	Oct. '72	£1·00
Part 5	Thyristors, thyristor stacks	May '72	£1·00
Part 6	Digital integrated circuits	July '72	£1·00
Part 7	Linear integrated circuits	July '72	70p
Book 2	**Valves and Tubes**		
Part 1*	Receiving valves, television picture tubes	Nov. '70	£1·00
Part 2*	Electro-optical devices, radiation detectors	Feb. '71	£1·00
Part 3*	Gasfilled tubes	Jan. '71	70p
Part 4	Transmitting and industrial heating valves	Oct. '72	£1·00
Part 5	Microwave tubes and components	Dec. '72	£1·00

		Publication Date	Price
Book 3	**Components, Materials and Assemblies**		
Part 1	Capacitors, resistors	May '72	70p
Part 2	Magnetic materials and components, piezoelectric ceramics	July, '72	70p
Part 3	Vinkor inductor cores	Dec. '72	£1·00
Part 4	RM inductor cores	Oct. '72	70p
Part 5	Radio, audio and television assemblies and modules	July '72	70p
Part 6	Circuit blocks, input and output devices, peripheral devices	May '72	70p

*Will not be revised in 1972.

**Numbers yet to be decided.

Note

Each of the eighteen parts can be bought separately. Orders (with remittances) should be sent to Central Technical Services, Mullard Ltd., New Road, Mitcham, Surrey CR4 4XY.